

Special Effects with a Galvanic Coil

WHEN LASER-LIGHT EFFECTS ARE WHAT YOU WANT TO PRODUCE, ANOTHER USEFUL COMPONENT TO HAVE IS A GALVANIC COIL OR GALVANOMETER. THE GALVANOMETER IS A DEVICE THAT IS USED TO MEASURE THE FLOW OF ELECTRIC CUR-

rent. You can make a simple one by winding 40 or 50 turns of magnet wire around a compass, in a north/south direction. When a voltage is applied to the coil, the compass pointer will swing from the north position to either east or west.

The same effect can be obtained by placing an external coil next to the compass at the east or west position. An applied voltage will then pull the compass pointer to the coil. You can use this principle to create an oscillation effect, which can be used effectively in light shows; in fact, it is the basis for many of the commercial light-show machines. The principle is simple: Feed a modulated voltage to the galvanometer coil, and the pointer movement will follow the signal or data being introduced in much the same way the laser flickers from electronic modulation.

Building a Galvanometer

For our purposes, a compass is far too delicate to be of much use. The alternative is to build a galvanometer that has a more rugged pointer—one that will support a mirror. There are a couple of ways to do that. But first, we must remember that the compass pointer reacts to the electromagnet, or coil, because it is also magnetized. So the first step is to rig a magnetic field on a vertical shaft that will pivot back and forth. That field can be permanent or electric, depending on the need; the merits of each will be discussed later.

Next, an electromagnet has to be positioned perpendicular to the vertical field, and at a distance that will allow the first magnet to clear the second in a full swing. With both components in place, a current sent through the electromagnet will attract the opposite pole of the vertical magnet and bring it around. I know that this is a rather simplistic explanation, yet it is the physical principle that makes the galvanometer function, and makes it a useful tool for visual productions.

WARNING!!!

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As mentioned before, the vertical field magnet could be either permanent or electric. A permanent magnet offers

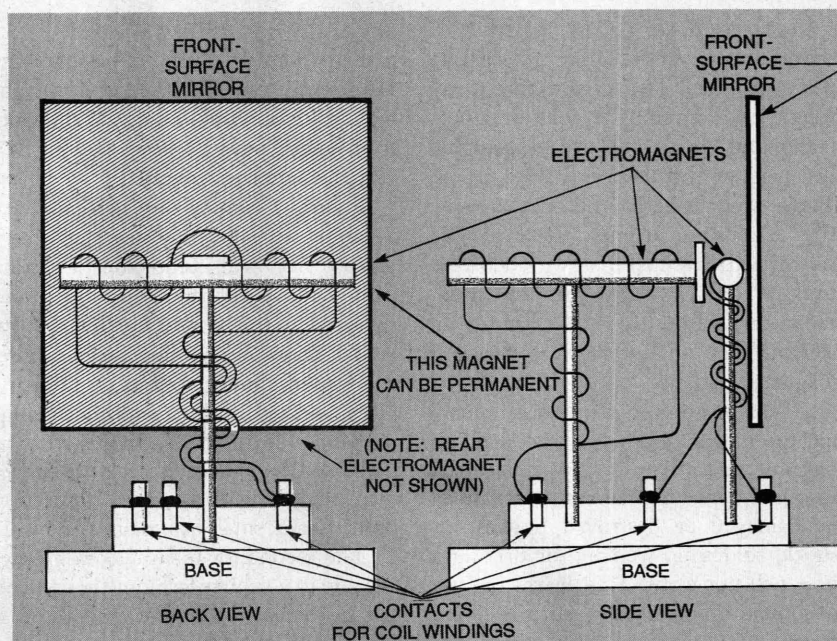


FIG. 1—A DETAILED LOOK AT A HOME-MADE galvanic-coil mirror assembly. It provides a great way to add a variety of movements to your light shows.

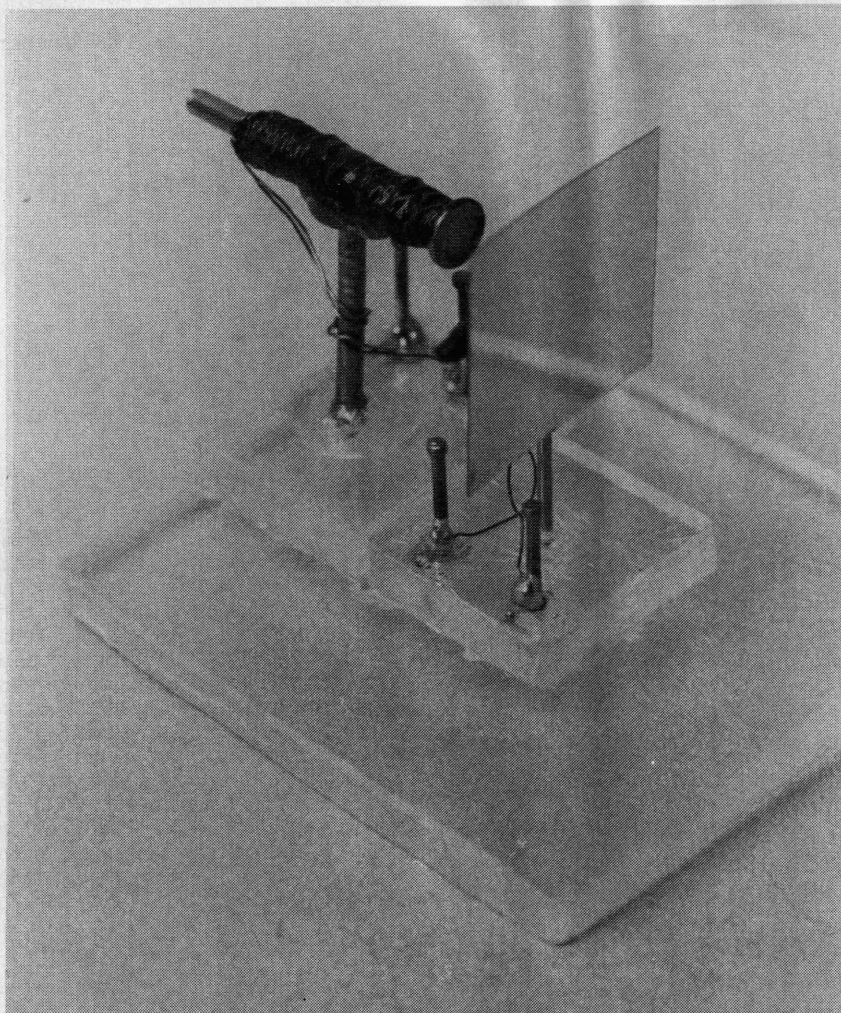


FIG. 2—THIS PHOTOGRAPH SHOWS the completed assembly. Note that the magnet-wire coils should be coated with glue or liquid rubber to hold them in place.

the advantage of not requiring a separate power source. The disadvantages are weight and mounting. If you use a permanent magnet, it will have to be a small bar magnet that is reasonably strong. Those tend to be a little on the heavy side. There are magnets that resemble rubberized strip material, with adhesive backing, that could also be used. However, while they are lighter in weight, they do not have as strong a magnetic field.

The next point to consider is mounting the magnet to the vertical shaft. In our unit, that will be a standard flat-head iron nail. Since permanent magnets can be damaged or destroyed by heat or shock, soldering, welding, or drilling a hole will not work. That leaves the use of some form of adhesive, such as epoxy, super glue, hot glue, or whatever else works for mounting.

The other alternative is an electro-magnet. The power requirement is the

primary disadvantage of using one. However, it can be turned to an advantage, since we could modulate the electromagnet's power source and further vary the patterns generated.

Figure 1 shows our galvanometer assembly. To build your own, start with a flat-top iron nail, and solder a second iron nail, with the top and point removed, to the head of the first to form a "T" shape. Using fine magnet wire, wind several hundred turns on the cross piece, and secure them with glue or liquid plastic. Bring the lead wires down to the base, leaving some slack for movement, and solder them to the connection posts. Again, small iron nails work well.

The second, or main, electromagnet is made in a similar fashion (though you can leave the head on the cross piece, if you wish) with 200 to 300 turns of wire on an iron nail. The coil is fixed in place with glue or plastic. The base that the magnets are attached to is made of wood

or Plexiglas. A front-surface mirror is then attached to the vertical, or moving magnet, and the galvanic coil is ready for use. The photo in Fig. 2 shows the completed assembly.

In this configuration, the mirror will move on the X axis. To get Y-axis movement, merely turn the assembly on its side. Most of the commercial units, called scanners, use two coils, one for each axis. When connected to control circuitry, they develop an expansive variety of visual images.

Using the Galvanometer

Despite its relative simplicity, you can create a wide variety of visual effects with the equipment you have just made. The control circuit could be any audio source, including a portable-amp/microphone arrangement, or a pulse source for an established pattern. Figure 3 shows a simple pulse-generator circuit that could be used for that purpose.

When the signal from the control circuit is applied to the main electromagnet, the vertical field will be activated proportionally. That moves the mirror and creates the oscillation effect.

You could also use the galvanometer with one of the optical-table assemblies we built in the previous months. For example, Fig. 4 shows one such modification—in this case from the December, 1996 column—with our galvanometer used in place of a speaker modulator. The reflected image that appears on the screen should be a combination of that produced by the revolving mirrors and the oscillating pattern generated by the coil. The dual revolving mirrors could also be replaced with additional galvanometers. If you incorporate beam splitters, a single laser can supply the light for various galvanic coils to produce a maze of intricate images. At the risk of sounding redundant, the only limit to a dramatic world of visual imagery is your imagination.

Once the combinations have been tested and a favorite light-table assembly has been selected, try introducing special effects to the presentation. Colloidal suspensions such as smoke, fog, dry ice in water, or fine powders puffed into the air all produce a reflective surface for the laser beam and make it visible when it passes through them. By introducing the suspensions at strategic locations, you can add another dimension to the show. Reflecting the beam off the surface of moving liquids provides an excellent

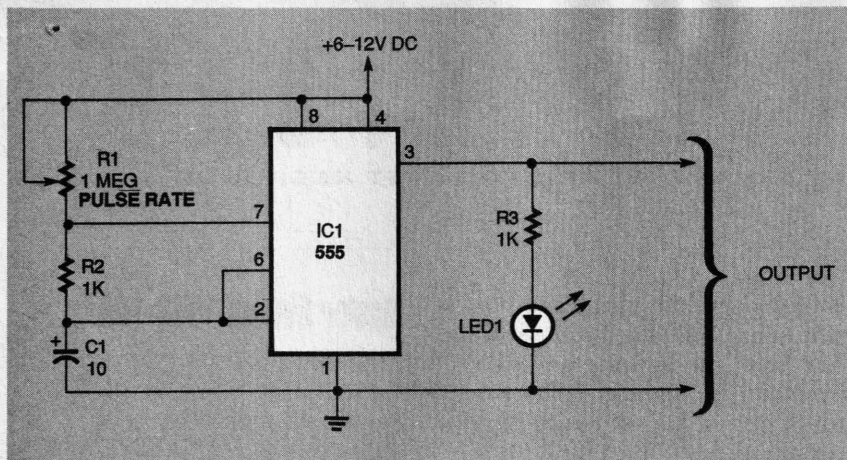


FIG. 3—THIS PULSE GENERATOR circuit could be used to drive the galvanic-coil/mirror assembly. Resistor R1 controls the pulse rate.

source of unexpected pattern variations. Try Mercury metal, if available, but be careful; it has a very high surface tension

and will get away quite easily. That's why they call it quicksilver!

Warping the reflective surfaces is

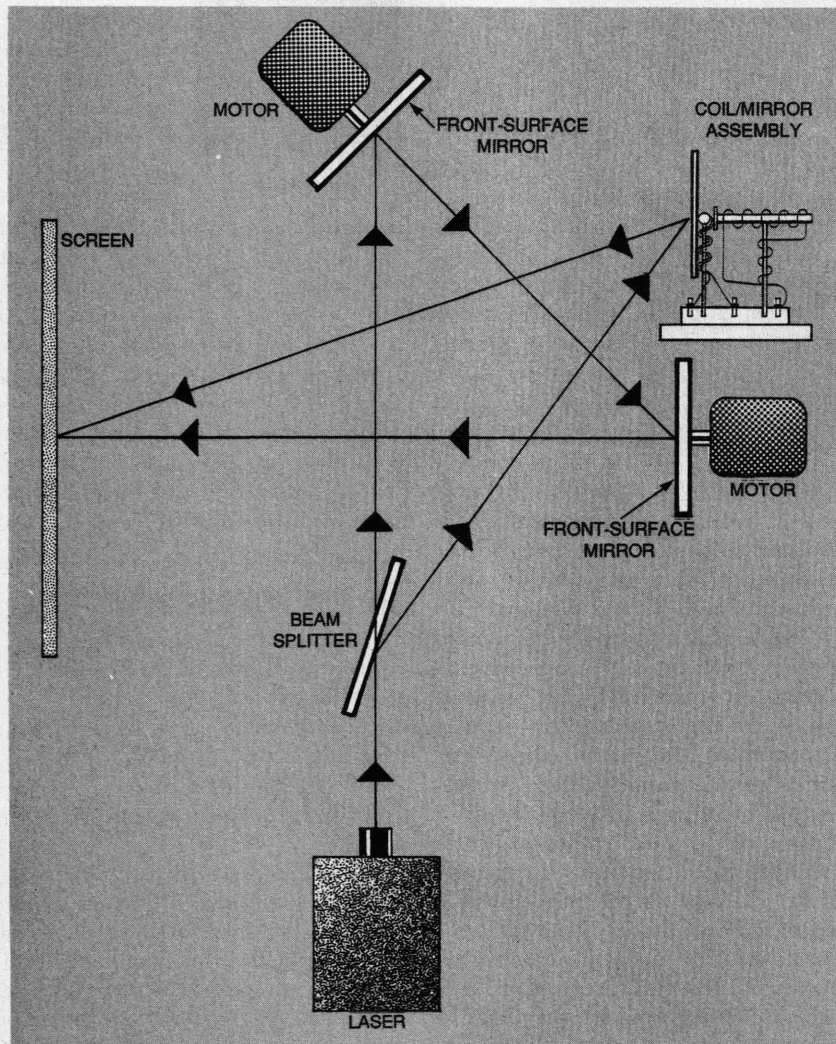


FIG. 4—HERE, THE GALVANIC-COIL ASSEMBLY has been combined with rotating mirrors and beam splitters to produce an intricate display. Note that the basic set-up was first presented in the December, 1996 issue; here a speaker modulator has been replaced with the galvanic assembly

another method of obtaining weird fun-house effects that add spice to the production. Aluminized Mylar materials are especially suited to that application. High sensitivity luminous paints or tape can hold a trace of the beam for a short period of time and give some outstanding results. These are just a few of the ideas that come to mind at the mention of special effects. With some thought and general observation, many more will present themselves.

If you have some ideas of your own that you would like to share with other readers, please put them down on paper and send them to Laser Editor, **Electronics Now Magazine**, 500 Bi-County Blvd., Farmingdale, NY 11735, or via e-mail to lartronics@aol.com.

That wraps things up for this issue. Next time we will take a look at how you might use your laser to conduct scientific experiments or as the basis of fascinating science-fair projects.

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ELECTRONIC GAMES

BP69—A number of interesting electronic game projects using IC's are presented. Includes 19 different projects ranging from a simple coin flipper, to a competitive reaction game, to electronic roulette, a combination lock game, a game timer and more. To order BP69 send **\$8.00 (includes s&h)** in the US and Canada to **Electronic Technology Today Inc., P.O. Box 240, Massapequa Park, NY 11762-0240**. US funds only. Use US bank check or International Money Order. Allow 6-8 weeks for delivery. MA07



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Camcorder Zoom-Lens Systems

A ZOOM LENS IS ACTUALLY A SYSTEM OF SEVERAL LENSES THAT ARE COMBINED TO PROVIDE VARIABLE MAGNIFICATION OF THE PICTURE WHILE SIMULTANEOUSLY KEEPING THE IMAGE SHARPLY FOCUSED. WHEN THOSE SYSTEMS FAIL AND

require troubleshooting, you really need to know what you are doing. That requires a basic understanding of how each of the individual components that comprise the lens system operate, as well as how all those components interact.

There are four basic elements in every modern zoom-lens system. They are: the lens assembly, the zoom control, the focus control, and the iris control. We will look carefully at each of those elements; once we understand how and what they do, we will go on to troubleshooting techniques.

Basic Lens Assembly

The zoom-lens system is designed so that errors in magnification and focus are minimized whenever the system moves back and forth (zooms in or out). That is especially important at the beginning and ending positions of the zooming process. It is also important to keep those errors from becoming excessively large at any of the intermediate positions.

A lens system that is not very critical of the focus throughout the zoom range is called an optically compensated *varifocal* system. Figure 1 illustrates such a system. All three operating positions are shown in that figure. The focal length of the lens is determined by the camera operator who moves a zoom handle forward or backward on the camcorder. That can be either a manual or a powered

adjustment. Either way, the lens elements that produce the zoom action are stationary in the lens cylinder. They move as a group, not individually in relationship to each other.

Some of the early variable focal-length lenses contained as many as 15 or more separate groups of elements. Modern camcorders have as few as four. As the number of element groups has been reduced, the complexity of the electronics required to maintain proper zoom/focus characteristics has increased. The optical zoom ratio (the ratio of maximum to minimum power) is, in general, 6:1. But in some designs that range is as large as 15:1 or more.

The lens shown in Fig. 2 is typical of

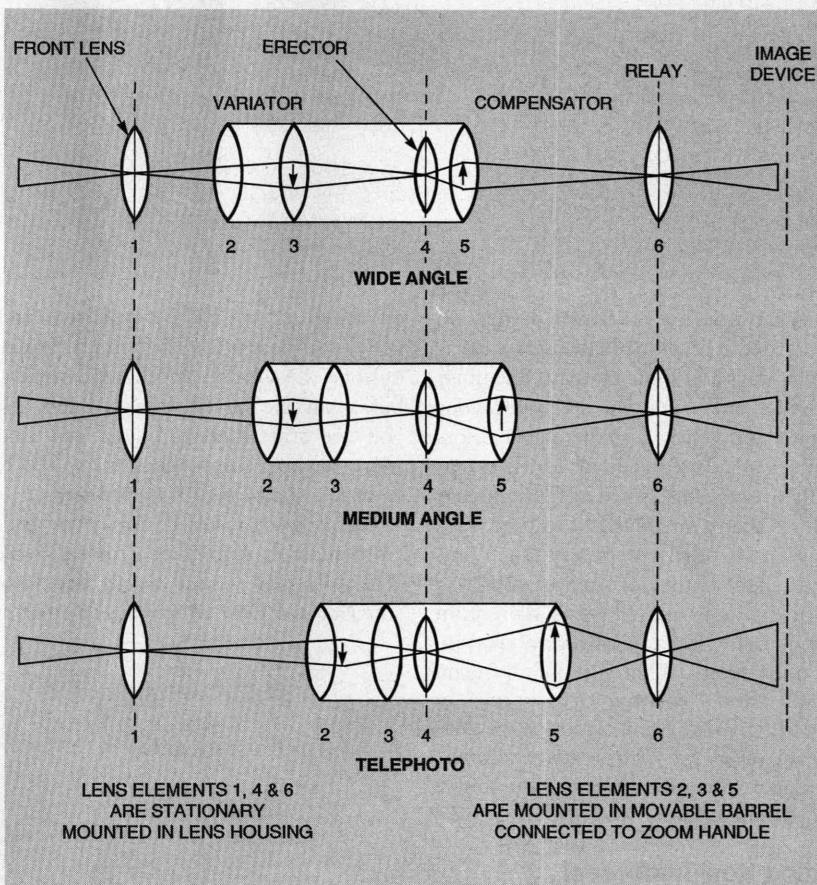


FIG. 1—A SIMPLE SLIDE ZOOM LENS SYSTEM that was used in early video camcorders. The lens elements are shown in all three common positions.

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Hitachi Home Electronics (America), Inc.
Suwanee, GA 30174

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Brodhead Garrett

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(800) 321-6730

Castcraft

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Memphis, TN 38187
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When you do your own design, do not forget to provide reliable safety isolation, always add some effective line noise filtering and shielding, and treat your input audio gently. Use the purest colors you can get. Avoid yellow because it typically overwhelms. Since an incandescent lamp is pretty much yellow to start with, filtering for other colors very much limits the light output. Go out of your way to prevent any and all white light from any pinholes, leaks, color recombination, or poor filters. There's a lot more on lamps and lighting resources in MUSE95.PDF. Available on www.tinaja.com or in the *Tech Musings* reprints. Grainger, of course, is your prime source for lighting options.

Going Non-Traditional

What about non-traditional color organ architectures? LEDs are one possibility, but power levels may be restrict-

ed, the blues will still be expensive, and you'd still be stuck with limited display options. Lasers are another possibility, but again the blues are a problem. (If you are still interested, you can find more on laser approaches from MWK Industries or Meredith Instruments.)

What if you instead combined a spectrum analyzer chip with a screen saver? The ultimate would be to use very sophisticated DSP techniques to attempt

to isolate music instruments. Let each instrument be a color blob on the screen. Some kind of tracking comb filter, maybe. The louder the instrument, the bigger and brighter the chosen blob.

Those blobs could wander around lava lamp style, continually showing new and interesting variations. Different algorithms could also be keyed to mood or music style. Texas Instruments has a new DSP demo board with a full real-time audio spectrum analyzer built in. It might make an interesting front end for audio-to-color displays.

New Tech Lit

The sudden demise of traditional data books might soon be upon us. That's because they have become outrageously expensive to produce and maintain. Instead they are being replaced by electronic versions. For example, new and free Acrobat-based CD-ROM versions of full-product data are freshly issued by Texas Instruments in a *Designer's Guide and Data Book*, and by Linear Technology in a similar publication.

A great source for the tech-education stuff is Brodhead-Garrett, who publish a free catalog. They stock everything from solar energy to injection molds.

Rex Supply prints a fat catalog chock full of machine-shop stuff. Stemgas Publishing has magazines and reprints of interest to collectors of antique gas or steam engines.

The Boy Mechanic III is the latest reissue from Lindsay Publications. It is a thick 1919 reprint full of wondrously essential projects such as full details on building your own cement kiln.

From Tapestry comes a complete line of fancier ink-jet media—glossies, clear, cling, golds, silvers, etc. Ace Plastics has a wide variety of unusual plastic shapes including tubing, rod, squares, rounds, balls, cubes, rings, diamonds, wedges, and more. From Castcraft comes a new flyer on casting and mold-making.

Contract Professional is a brand new magazine for independent tech consultants. If you are interested in that, you should also see the *Synergetics Consultant's Network* you'll find on www.tinaja.com.

I've still got some great surplus buys on Tektronics 1230 and Fluke 9010 logic analyzers. Most of the mentioned resources appear in our Names and Numbers or AC Phase-Control sidebars. Be sure to check there before calling our no charge U.S. help-line or visiting my www.tinaja.com Web site. **EN**

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the lens system in many modern camcorders. It uses a device called a complicated cam to move the lens elements to produce the necessary zoom ratio. The complicated cam is a cylinder that has machined slots to position the lens elements in varying degrees of zoom or macro. (Macro is the opposite of zoom; it is used to focus on an object that is in close proximity to the lens.) The relationship of the two lens elements to each other changes according to the zoom and macro ranges. Also, the lens elements shift in direction depending on whether they are zooming in or zooming out.

The rear zoom element comes further forward during macro than it does during zoom. That is what moves the focal distance closer to the front lens. The precision of the lens group's movement is controlled by the complicated cam. To compensate for focus errors during the zoom process, an adjustable back-focus lens is located in the rear of the lens assembly. It is adjusted to eliminate errors throughout the full zooming range of the lens and to maintain image focus on the pickup device. No electronics are required to monitor movement of the lens-element groups.

In older camcorders, zoom controls were often simple electric motors that were fed a reversible voltage to change the rotational direction of the zoom barrel connected to the complicated cam. The zoom and focus functions were separate in those types of lenses.

How Focus Works

To produce sharp, high-resolution pictures, the camera lens must have good optical qualities. Equally important, the image that the lenses produce must be accurately focused on the sensor. The final element that determines picture resolution is the image sensor itself. It takes a high-quality lens coupled to a high-quality image sensor to produce a high-resolution picture.

Focusing is handled by moving the lens closer to or further away from the image sensor. The lens is closest to the sensor when it is focused on distant objects (such as landscapes), which are said to be at optical infinity. For most normal-focal-length camcorder lenses, infinity means anything further than 30 feet (10 meters) away. Focusing on subjects closer than infinity demands moving the lens away from the sensor. That is done with either a helical-thread lens (that can be operated manually or be driven

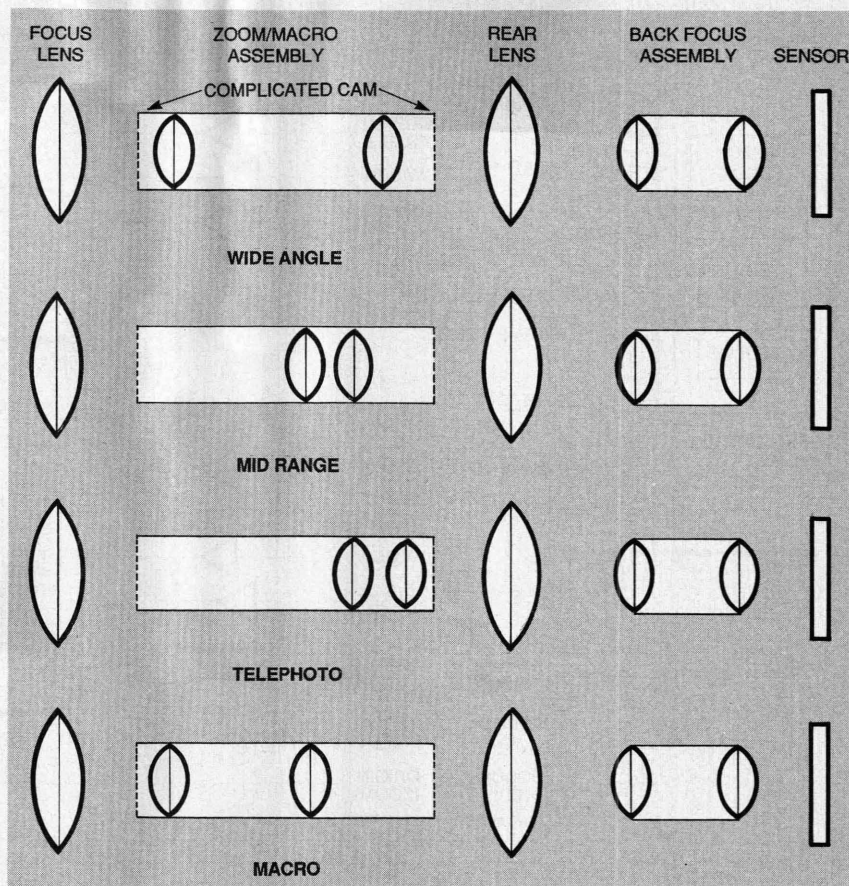


FIG. 2—MODERN CAMCORDER LENSES consist of fewer elements. Here's a look at how the elements appear in the four most commonly used positions.

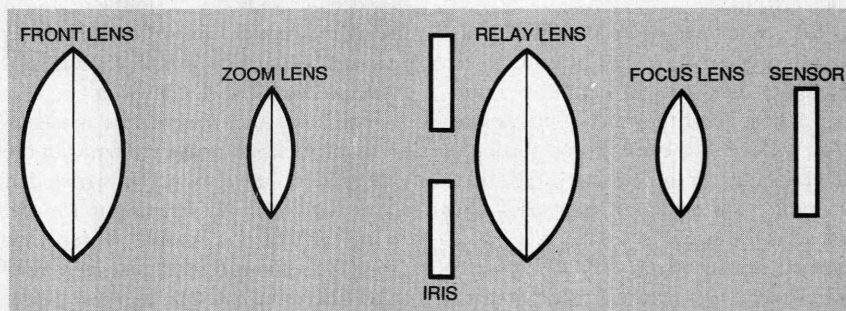


FIG. 3—THIS FOUR-ELEMENT INNER FOCUS LENS is used in many current Hitachi camcorders. Similar systems can be found in camcorders from other manufacturers.

ven by a gear-reduced miniature DC motor), or by an internal sliding lens assembly (that is driven by either a linear-drive or a stepping motor).

Focusing is very straightforward with manual-focus lenses. The helical-thread lens is rotated (manually or electronically) until the image in the viewfinder is sharp. That method leaves two variables: Is the viewfinder properly focused? Is the operator's eye properly focused to the viewfinder? An auto-focus system, on the other hand, removes the viewfinder from the focus system, leaving the

alignment of the auto-focus system as the only remaining variable.

Figure 3 is a block diagram of the lens system that is used in many current Hitachi camcorders, and similar systems are found in many other brands of camcorders. The lens consists of only four elements and has a stationary front-lens element. The focus-lens element is located at the rear of the lens assembly, and functions as a back-focus element. The zoom lens has been reduced to a single lens-element group, and the complicated cam has been eliminated. Many lens-con-

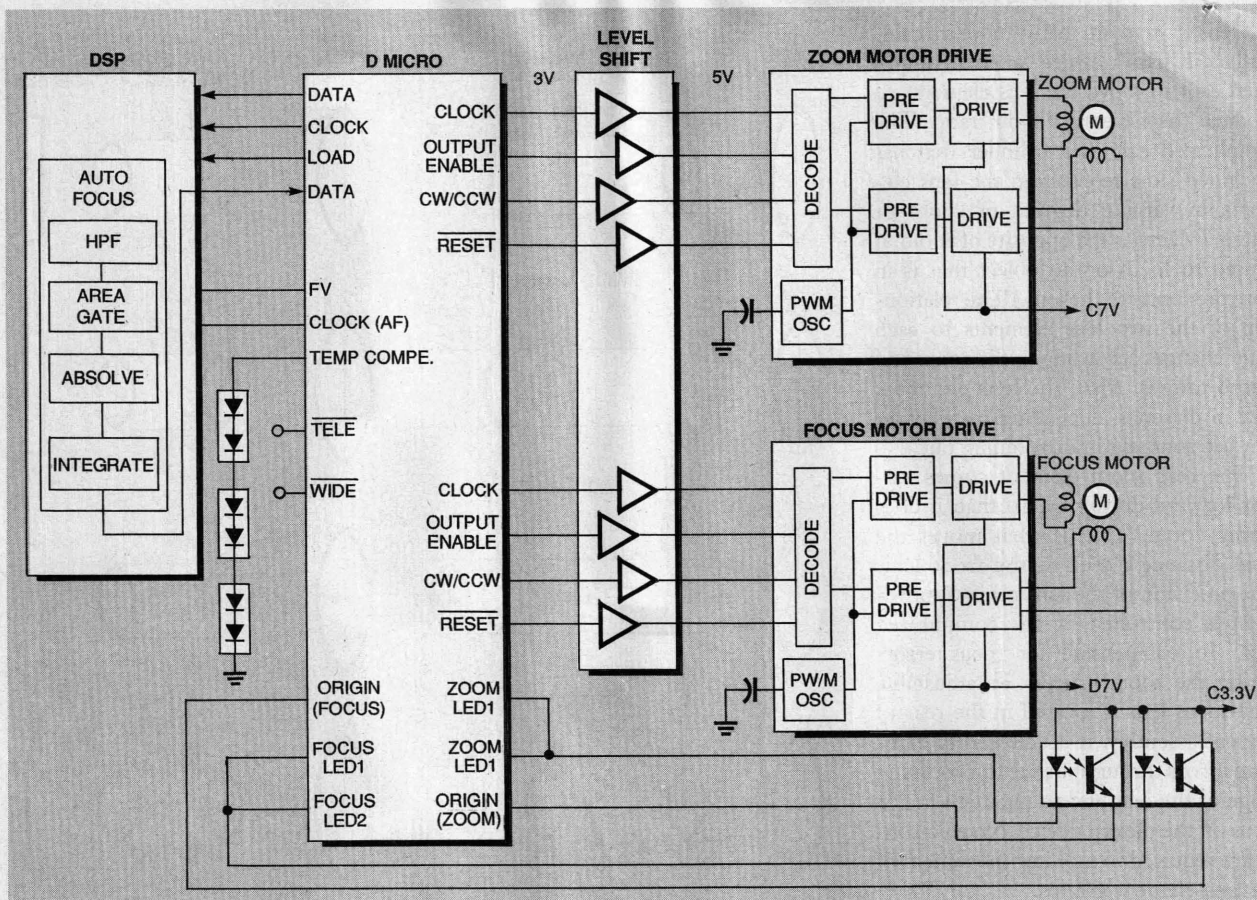


FIG. 4—A BLOCK DIAGRAM of the zoom/focus drive system. Note that zoom control circuitry is virtually identical to the focus control circuitry.

control functions are now provided entirely by electronics.

The system relies on the actual video signal processed by the camcorder for focusing. As shown in Fig. 4, the luminance portion of the video signal is sampled and fed to the auto-focus portion of the digital-signal-processing (DSP) IC as an evaluation signal. A high-pass filter converts the signal to a voltage. The frequency characteristics of the luminance signal increase in proportion to the sharpness of the focused image, and the developed voltage increases accordingly—the higher the voltage, the sharper the luminance signal and the better the focus.

That signal goes through its own signal processor to provide information to the digital microprocessor, which evaluates the data and determines whether or not the image is in focus. If it is not, it outputs motor-control data to the drive circuit to move the focus motor in the appropriate direction to bring the image into focus. The control data consists of four elements: the output-enable signal, the direction of rotation signal (C/CCW), the clock signal, and the reset signal.

The focus-drive motor is a small step-

ping motor driven by a 6-volt pulse from the focus-drive IC to rotate the motor in the right direction. The focus-drive IC is powered by the C7V power bus originating from the DC-DC converter.

The zoom-control circuitry is identical to the focus-control circuitry, with the exception of the evaluation signal that starts the process. The input for the zoom control is a simple switch that grounds an input to the microprocessor to tell the zoom the direction in which it should go (telephoto/wide). That information is compiled by the microprocessor and is output to the motor drive IC to rotate the zoom motor in the proper direction.

Since there is no back-focus assembly, the focus and zoom circuits must constantly interact to compensate for lens irregularities as well as normal focusing. That information is calibrated by the focus portion of the alignment procedure.

The first step in adjusting auto focus is the zoom/focus tracking. The lens is automatically run through the full zoom range, and focus tracks along with zoom to eliminate any lens errors. When this

process is finished, the information is stored and recalled to use as the zoom lens is moved. This means that the focus lens has become a back focus lens during the zooming process.

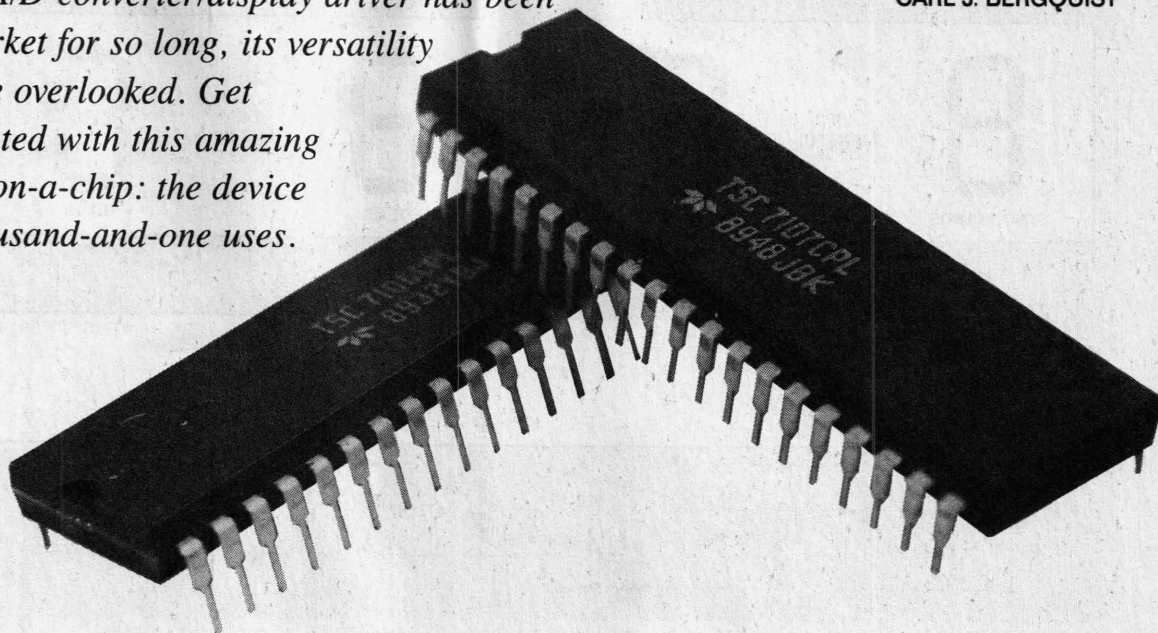
Now you know just about everything there is to tell about how the zoom-lens system works. Next time we will look at the troubleshooting procedures you will need to successfully service camcorder zoom-lens systems.

EN



The 7107 A/D converter/display driver has been on the market for so long, its versatility tends to be overlooked. Get re-acquainted with this amazing voltmeter-on-a-chip: the device with a thousand-and-one uses.

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USING THE 7107

With the invention and introduction of the seven-segment light-emitting diode (LED) display, adding an information display to the front panel of almost any project has become commonplace. Digits are easier to read than analog meters, and tend to "dress up" the appearance of the completed unit.

Those displays, of course, do not light themselves, but rely on logic and driver circuitry to display the desired information. The combination of a binary or decade counter with a decoder/driver is one way to display information on the LEDs, especially in clock applications, but a more versatile method is an analog-to-digital converter.

A number of options are available along that line. Combining a Motorola 14433 3½-digit A/D converter or an RCA CA3162 A/D converter with an RCA CA3161 BCD-to-7-segment decoder/driver chip is one way to go, but for ease of use, low cost, and low component count, it's hard to beat the Intersil/Harris ICL7107/7106 3½-digit A/D converter/display driver.

Originally designed primarily as a voltmeter, the 7107 is a combination analog-to-digital converter, reference voltage, timing clock, and decoder/driver circuit all on a single CMOS IC. It features high accuracy, low power consumption, and directly

drives instrument-sized, common-anode, 7-segment LED displays. The 7106 is the liquid-crystal display (LCD) version of that device and incorporates a backplane driver.

The number of support components needed for operation of the 7107 are minimal: five capacitors, four resistors, a potentiometer, and the display. Add a split five-volt power supply, and the 7107 becomes a complete 3½-digit voltmeter with a full-scale reading of either 200 mV (199.9) or 2 volts (1.999), depending on the values selected for the support components. Another attractive feature is that since both chips have been on the market for several years, their price through catalog distributors has dropped to below four dollars each.

With the basic voltmeter circuit, any measurement, value, or quantity that either exists as a voltage or can be converted to a voltage will provide the input. The 7107 is then easily calibrated to display that measurement or value digitally.

So, the possibilities are limitless. Volt-

age, current, resistance, inductance, light, temperature, speed, capacitance, frequency, etc., can all be directly displayed through simple transducer circuits. We'll show you how to do that, and armed with that knowledge, it is very convenient to either equip projects and new designs with digital displays, or modernize older gear.

7107 Circuit Operation. Figure 1 is the schematic for a basic 2-volt configuration. Power for the chip is applied to pin 21 for ground, pin 26 for -5 volts, and pin 1 for +5 volts. Pins 2-20, and 22-25 connect the device to the LED display, with pin 19 driving the ½ digit, or the leading "1", on the display. Pin 20 indicates when the input voltage drops below 0 volts, so it can be used to drive a minus sign on the display.

Pins 30 and 31 accept the negative and positive inputs, respectively, while pins 35 and 36 handle the reference voltage generated by a resistor-potentiometer divider network. The 10

smaller value also improves recovery speed of the circuit. A value of 1,000 ohms for potentiometer R3 in the lower-voltage meter is increased to 25,000 ohms to provide a wider reference range for the higher-voltage meter.

For maximum accuracy and stability, it is best that all capacitors, with the exception of C5, be of the Mylar variety. With its lower dielectric absorption, that type of capacitor will provide superior performance. Capacitor C5 can be a standard ceramic disc or silver-mica type, and all resistors are 5% carbon film.

One disadvantage of the 7107 is the need for a split 5-volt power supply, but there are several simple ways to meet that requirement. If a bench supply is at hand, it usually has ± 5 volts available, either as a separate output or through the use of the adjustable supplies. If that source is not convenient, or a separate self-contained supply is preferred, Fig. 2 shows three different approaches.

In Fig. 2-a, input power from a small step-down transformer is first rectified by bridge rectifier BR1 and smoothed by capacitor C1. A pair of 1N4733 Zener diodes (D1 and D2) regulate the input power to the needed 5 volts.

PARTS LIST FOR THE 7107 VOLTMETER

SEMICONDUCTORS

IC1—ICL7107CPL A/D converter/LED display driver, integrated circuit
DISP1, DISP2—dual 7-segment common-anode LED display (MAN6710 or similar)

RESISTORS

(All fixed resistors are $\frac{1}{4}$ -watt, 5% units.)

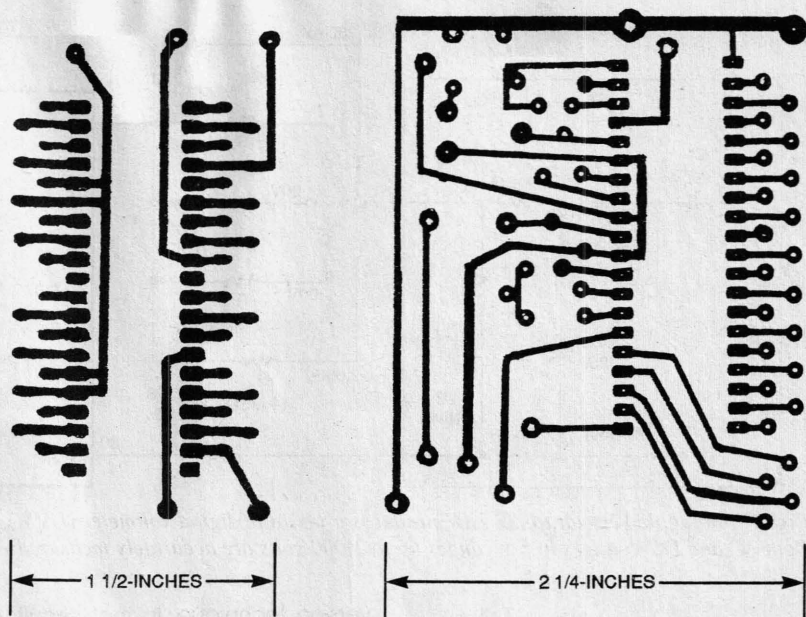
R1—47,000-ohm/470,000-ohm
R2—1-megohm
R3—1000-ohm/25,000-ohm, potentiometer, (see text)
R4—22,000-ohm
R5—100,000-ohm

CAPACITORS

C1—0.22- μ F, Mylar
C2—0.47- μ F/0.047- μ F, Mylar (see text)
C3—0.01- μ F, Mylar
C4—0.1- μ F, Mylar
C5—100-pF, ceramic-disc

ADDITIONAL PARTS AND MATERIALS

IC sockets, hookup wire, etc.



Here are the foil patterns for the basic 7107 voltmeter and the LED display. A split power supply and a handful of passive components turn the 7107 into an accurate, easy-to-use instrument. Putting the display on a separate board makes the project more versatile in terms of mounting arrangements. You can also combine both boards into one if you need to.

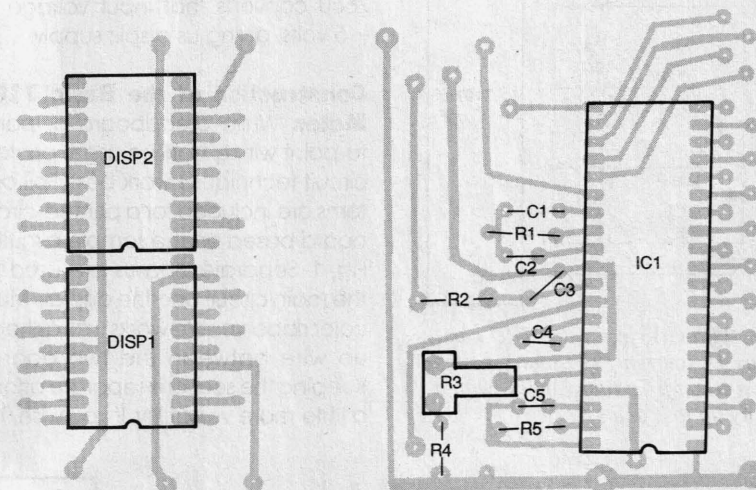


Fig. 3. If you want to build your own 7107 voltmeter, you can follow the parts placement diagrams here. Using sockets for the 7107 and the displays make troubleshooting easy.

Actually, the diodes are rated at 5.1 volts, which is well within the 7107's tolerance. Resistors R1 and R2 limit the current that flows through the Zener diodes themselves. The remaining circuitry is a standard "rail" power supply.

Figure 2-b illustrates a similar configuration utilizing positive- and negative-output linear voltage regulator ICs. However, we are using a center-tapped transformer to form a true split power supply. A disadvantage of a center-tapped transformer is the added expense. However, never try to

get around that by using a non-center-tapped transformer and connecting one side of the AC input to ground. Although that would work, it would create a "hot" ground: a potentially dangerous situation. Any extra money spent on the proper transformer will be far cheaper than spending several days in a hospital should some component in the circuit fail. Be sure both the 7805 and 7905 voltage regulators have adequate heat sinks.

Figure 2-c uses a voltage-con-

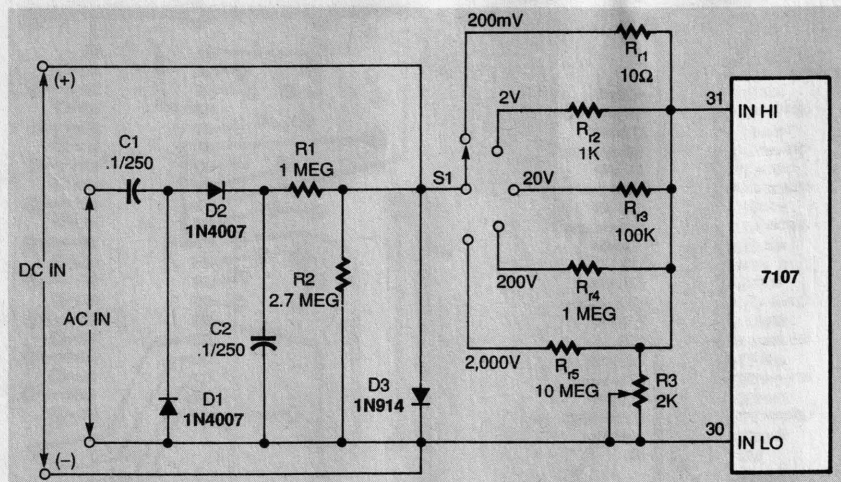


Fig. 4. Adding this circuit to the 7107 produces a versatile digital voltmeter (DVM). Both AC and DC voltages in five ranges up to 2,000 volts are accurately measured.

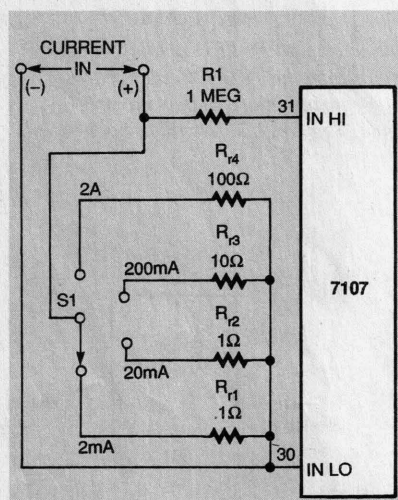


Fig. 5. With this circuit, the 7107 can measure DC current. The rotary switch lets you select from four ranges, with a capacity up to 2 amps.

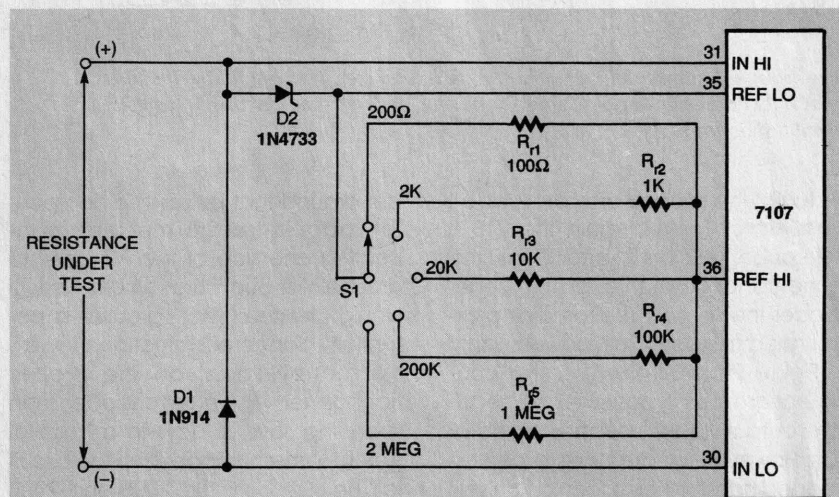


Fig. 6. Although you could measure resistance with a simple voltage-divider circuit, the use of a Zener diode and a switch-selectable resistor to change the 7107's reference voltage makes it easier to measure resistances up to 2 megohms.

LED readouts and main circuit can be combined onto one board. It's a simple matter to merge the two patterns together.

Regardless of how you handle your PC boards, use a 40 pin IC sockets for both the 7107 and the LED displays. That makes changing the 7107 or the displays easy if you need to. The parts-placement diagram in Fig. 3 will aid in component insertion at the proper locations. When installing the discrete components, the only orientation to observe is the IC itself, as the capacitors are non-polarized.

Like the display circuit, the power-supply circuit can either be put on the main PC board or kept separate. Convenience or personal preference will, of course, influence which way you decide to go. With all three sections completed and connected together, we can move on to the next step.

Testing and Operation of the 7107

Meter. Before installing the 7107 chip itself, confirm that the correct voltages are present at the power supply, and at pins 1, 21, and 26 of the 7107 socket. If a problem is encountered, double-check the wiring and component placement for errors. Once the proper voltages are verified, disconnect the power and plug the 7107 into its socket. Turn the power back on, and the display should read "000".

Apply +5-volts to pin 37, and "-1888" should be displayed. Pin 37 is the test input and activates all segments. A permanent switch can be added if the test function is to be used often. However, do not leave the test on for more than a few minutes if you're using the 7106. The test function on that chip applies straight DC instead of squarewaves to the LCD segments. That might cause the image to "burn" into the LCD.

Apply a known, stable voltage to the + and - inputs. Adjust potentiometer R3 until the display reads that voltage. The meter is now calibrated and ready to go.

Another method of calibration is to read the actual reference voltage at pins 35 (REF LO) and 36 (REF HI). For the 200-mV meter, that reading should be as close to 100 mV as possible. For the 2-volt meter, 1 volt is the needed measurement.

That procedure will also be useful if

an abnormal voltage reference is needed. In some applications where the transducer output is above or below the standard meter parameters, it is easier to adjust the reference voltage to accommodate the transducer, rather than attempt to correct the transducer's output voltage. In any event, always adjust the reference to one half the input voltage.

Additionally, a decimal point can be displayed by hard wiring a 470-ohm resistor from ground to the pole of a four-position rotary switch. With the four contacts connected to the four decimal points on the display, the decimal point can easily be placed wherever it is needed.

Applications and Input Transducers.

With the basic digital meter built and tested, it's time to put the unit to work. As we mentioned earlier, any value or measurement that is a voltage or can be converted to a voltage can be displayed, but only if we connect what we need to measure to the meter properly. That is where the input circuits and transducers come into play.

Voltage Measurements. Since the 7107 is a voltage-measuring device, let's look at that application first. Figure 4 is a circuit for a digital voltmeter (DVM) that will measure voltages from 0 to 2,000 volts in five ranges. The circuit can measure both alternating and direct current, with C1, C2, D1, D2, R1, and R2 acting as a rectifier/filter for AC measurement. Resistors R_x and R3 form a selectable input-voltage divider. Rotary switch S1 connects any one of the five resistors, R_1 – R_5 , for the selected voltage range.

Once that input circuit has been connected to the 7107, choose a known voltage source, either AC or DC. Select the appropriate range with S1, and adjust R3 to obtain a display that matches the known source. That will fine-tune the circuit for future voltage measurements.

Current Measurements. The next basic measurement to consider is current. We do that by passing the current through a shunt resistor. The shunt develops a voltage across itself in proportion to the amount of current flowing through it. That voltage can be measured by the 7107 meter.

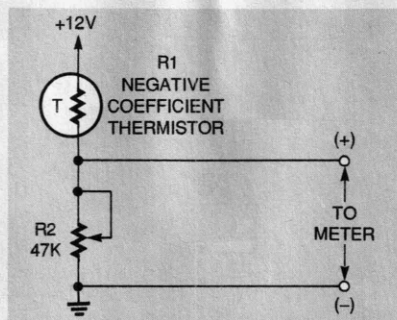


Fig. 7 A simple voltage-divider circuit using a negative-temperature-coefficient (NTC) thermistor is a low-cost and accurate method of measuring temperature with the 7107. An NTC thermistor's resistance drops as its temperature rises. If you switch the locations of R1 and R2, the reading on the 7107 will drop with rising temperature.

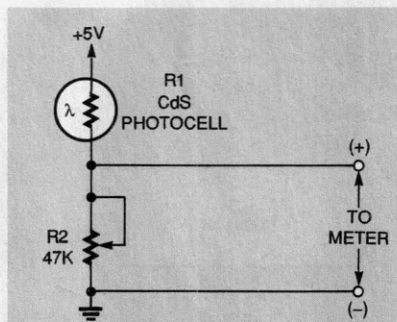


Fig. 8. We can measure light intensity the same way we measure temperature by substituting a cadmium-sulfide (CdS) photocell for the thermistor in Fig. 7. In most CdS photocells, resistance will drop with an increase in brightness. Should you need to have the 7107's display reading drop as the photocell gets brighter, or if you have a photocell that increases in resistance with exposure to light, simply switch the positions of R1 and R2.

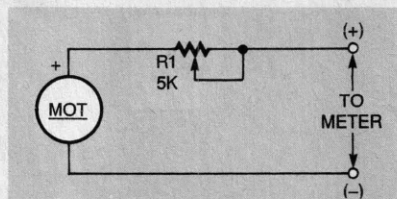


Fig. 9. You can use a small permanent-magnet motor as a tachometer. Attach the motor's armature to whatever spinning shaft you want to measure, and the voltage generated by the motor can be read directly by the 7107.

Figure 5 is the schematic of a four-range system, where R_1 – R_4 controls the range and acts as the shunt. Resistors R1 and R_1 – R_4 form a voltage divider that provides the input for the

meter. A rotary switch arrangement similar to the range selector in Fig. 4 works nicely for the ammeter as well. The values for the shunt resistor are selected by S1. Using 1% precision metal-film resistors gives the greatest accuracy.

Resistance Measurements. The last type of value commonly measured by most digital multimeters (DMMs) is resistance, and Fig. 6 illustrates a method for that. Zener diode D2 develops an external reference voltage, allowing S1 to select one of R_1 – R_5 to control the on-chip reference voltage. A DC voltage is generated across the unknown resistance, and that value is then processed by the 7107 and displayed.

That measurement could also be done with a standard voltage-divider circuit, but the problem with that method is that R_5 would have to be in the 500-megohm range for the 200,000- to 2-megohm setting. That is a difficult value to obtain, especially at precision levels. Adjusting the reference voltage results in a workable value for R_1 – R_5 .

Temperature Transducer. There are a number of ways to measure temperature. Figure 7 is a simple and surprisingly accurate method. In that configuration, a 47,000-ohm potentiometer provides as calibration, and a negative-temperature-coefficient (NTC) thermistor acts as the temperature probe. By applying 12 volts and setting the decimal point on the display between the units and tens digit, readings to an accuracy of $1/10^\circ$ can be achieved. Again, the circuit is a voltage divider.

Light Measurement Transducer.

The light transducer in Fig. 8 is almost identical to the temperature sensor. The only differences are a 5-volt supply voltage and the use of a cadmium-sulfide (CdS) photocell in place of the thermistor. Changes in ambient light shining on the photocell changes the resistance of the cell. That in turn changes the voltage produced by the divider. Any change in light intensity will be shown in the display. The display can be calibrated to indicate "f-stops" for photography work, or can be an arbitrary numerical value for which a chart or graph

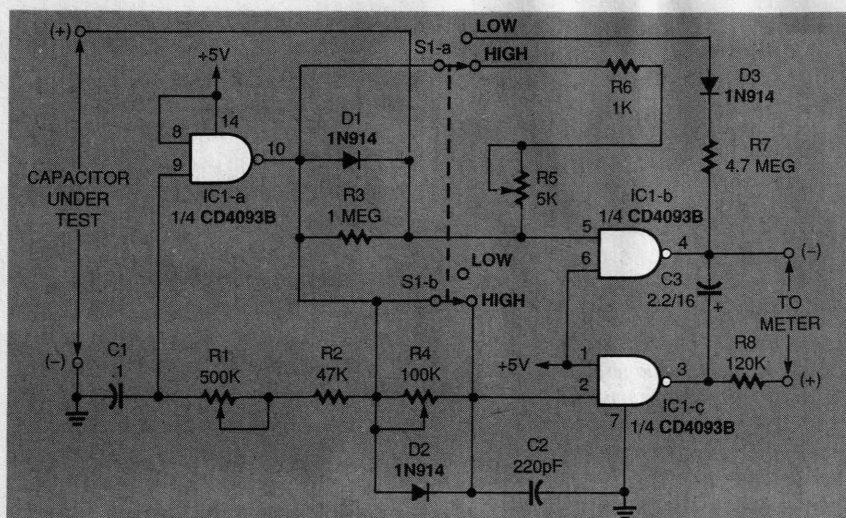


Fig. 10. With this circuit, the 7107 gains the ability to measure the value of capacitors from 2 pF to 2 μ F. Potentiometer R4 adjusts the circuit for a zero reading, while R1 and R5 calibrate the low and high ranges.

can be developed as appropriate. By using a photocell with a range of 500-ohms illuminated to 2- to 3-megohms dark, very sensitive results can be achieved with the circuit.

Those two circuits illustrate how any device that changes resistance due to any circumstance can be plugged into the divider network. You can try replacing the CdS photocell with a 1,000-ohm potentiometer. After calibrating with the 47,000-ohm variable resistor, the exact value of the potentiometer between 0 and 1,000 ohms will be displayed.

Speed Measurements. Figure 9 shows a method of measuring rotary speed. Since the 7107 is a voltmeter, and electromagnetic motors produce a voltage when their armatures are rotated, the measurement circuit is extremely simple. A 5,000-ohm potentiometer is used for calibration, with one end connected to the positive motor pole, and the other tied to the HI input (pin 31). The negative side of the motor goes to the LO input (pin 30). When the motor shaft is coupled to a rotating object, the voltage developed is proportional to the speed of rotation. That system can be used for tachometers, wind-speed indicators, generator-voltage monitors, and the like.

Capacitance Meter. One of the most useful bench tools around is the digital capacitance meter. Those meters will save not only time, but money as well. Surplus capacitors are

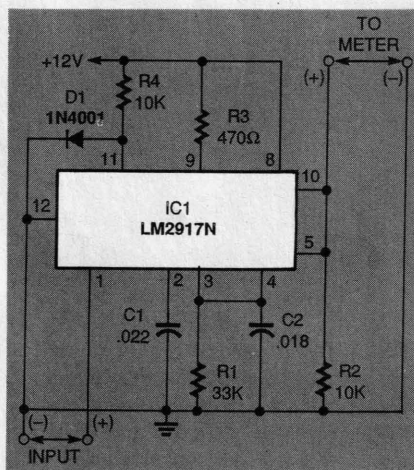


Fig. 11 Adding an LM2917 to the 7107's inputs lets you measure lower frequencies up to about 10 kHz.

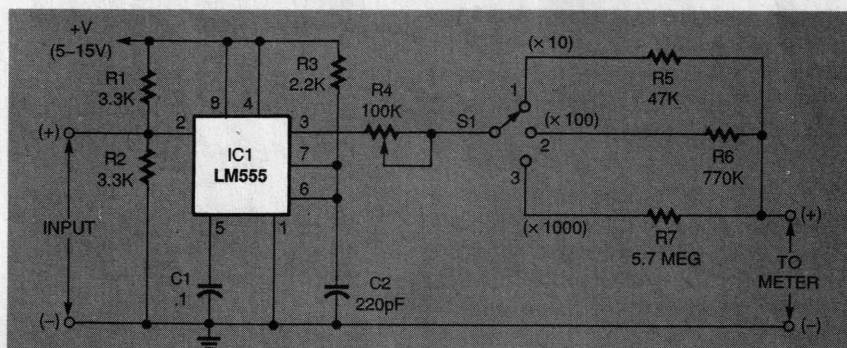


Fig. 12 To measure higher frequencies than the LM2917 can handle, this circuit does the job quite well. Frequencies in 3 ranges up to 2 MHz (1.999×1000) can be measured.

often sold by the pound at a greatly reduced price, but due to factory marking, or no marking at all, it can be hard to determine the values of those. A capacitance meter solves the problem, but most commercial meters are

relatively expensive.

Figure 10 is the schematic for a unit that makes use of a 4093 quad 2-input NAND Schmitt trigger, and will measure capacitors from 2 pF to 2 μ F in two ranges. The low setting handles 2 to 1,000 pF, while the high setting is for capacitor values from 0.001 to 2 μ F. The circuit hooks up directly to the 7107.

Once connected to the 7107, calibration is simple. Turn the unit on and adjust R4 for a zero reading. With S1 in the low setting, connect a 100-pF capacitor to the test leads, and adjust R1 for a reading of "0100". Then switch S1 to the high setting, and with a 1- μ F capacitor connected to the test leads, adjust R5 for a "1000" display. With that done, the meter is now ready to test and measure capacitors within the meter's range.

It is a good idea to conduct those last two steps with several known capacitors to fine-tune the calibration. While not shown in the schematic, it is common practice when using CMOS-technology ICs to connect the inputs of any unused logic gates (e.g., pins 11 and 12 of IC1) to either ground or the positive voltage. That will reduce power consumption and interference. On a final note, be sure the +5-volt supply is well regulated to maintain accuracy.

Frequency Measurements. Figure 11 uses the LM2917N frequency-to-voltage converter in a standard configuration. That device is a simple and

inexpensive way of displaying frequency with the 7107. The 2907/2917 series was originally designed for tachometer systems, and is limited to frequencies from DC to about 10,000
(Continued on page 79)

USING THE 7107

(Continued from page 60)

Hz. While it is not well-suited for frequency counters, it is an excellent choice for automobile tachometers, audio-frequency meters, pulse- and function-generator displays, and other such applications. The 2917 interfaces directly with the 7107, but might require a simple op-amp buffer at the input. If the output is erratic, a 0.1- μ F Mylar capacitor connected in series with the positive input will usually correct the problem.

For frequency counting, the LM555 timer is a better choice, as that chip will handle higher frequencies. Figure 12 is the schematic for a 555 frequency-to-voltage converter that can display up to a maximum of 1.999 MHz. The 100,000-ohm potentiometer (R4) functions as calibration, and S1 selects one of three resistors that determines range.

The display reading is multiplied by 10 for range 1, 100 for range 2, and 1,000 for range 3. A 1999 reading in range 2 would be equivalent to an input frequency of 199,900 Hz. A display of 0865 in range 3 would represent 865,000 Hz.

The range of the counter can be further expanded through the use of input pre-scalers. Those ICs will divide an input frequency, usually by 10, before applying it to the counter. A single pre-scaler would increase the input capability to 19,990,000 Hz, while cascading two pre-scalers would take it up to 199,900,000 Hz. Again, there is the multiplication factor to consider. When using pre-scalers, the displayed value is multiplied first by the range factor, then by the pre-scale factor to get the actual measurements.

The 11C90 and 95H90 ICs are two common pre-scalers, but the drawback with them is their expense. Those chips are not exactly economical, although the 95H90 can sometimes be found on the surplus market for as little as \$6 to \$10.

Other applications. Now that you're an expert on the 7107, the knowledge gained will perhaps prompt you to expand projects, designs, or equipment, and possibly inspire ideas for future projects. LED and LCD displays are highly informative, and can add

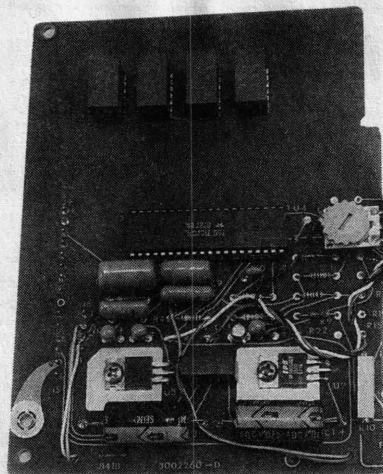


Fig. 13. A section of an alarm board, bought through surplus, that contains a 7107. This particular board also contains the $\pm$ 5-volt power supply.

to the functionality of many types of electronic equipment. A project that takes advantage of the 7107's voltage-measuring capabilities is the author's "Versatile Power Supply" article in the December, 1995 issue of **Electronics Now**. That project is a bench-type power supply with a 7107-driven LED readout that displays the level of the output voltage.

Only a few applications were covered in this article. With a little thought or as practical needs arise, many other ideas will present themselves. For example, the Intersil data book illustrates a circuit for a digital torque meter, using a resistance-dependent strain gauge.

One area to watch for an economical source of devices is the surplus market, as the 7107 and 7106 can be found on many circuit boards. The surplus circuit board in Fig. 13 is an example of that type of source. The chip is usually mounted in a socket, and can be removed for use in one of your own projects. Frequently, the basic meter circuit is intact, and can be used as is with minor modifications. The 7107 can be found in some alarms, satellite receivers, antenna-positioning units, salvaged digital panel meters, counters, controllers, and a variety of other equipment. Be sure and keep an eye out for them when surplus "bargains" are advertised.

So, pick up a couple of 7107s and try adding one to your next project. The chip is quite easy to work with, and the results you could achieve can be very gratifying. Ω

MILLI-OHM ADAPTER

(Continued from page 42)

Run a wire directly from the ground connection of C1 on the board to a lug on the 10-32 stud supporting the swing arm. The common connection including C10, C11, C12, R9, and J1 is the negative Kelvin lead, which is *not* to be connected to ground on the board. Connect the Kelvin lead in this fashion: Run an insulated lead from the common connection of C10, C11, and C12 through a grommet located in the top of the cabinet near the stud, then through an "eyelet" located on the stud itself, and finally make the connection to the lug on BP2.

Mount the completed board on a pair of 1/2-inch spacers in the bottom of the cabinet and connect the remainder of the panel components to the board. Install the ICs in their sockets and close the cabinet. Prepare an output cable using a twisted pair of wires (do *not* use coaxial cable) with an RCA phono plug on one end, and a suitable plug for your DVM on the other end. A 12-inch length is sufficient.

Calibration. Power up the adapter and allow it to warm up for about five minutes, allowing the temperature to stabilize in the cabinet. Connect the adapter to your DVM, open the cabinet, and connect a clip-lead jumper from the base lead of Q1 to its emitter lead. The resulting voltage, which can now be read on your DVM, is the *offset voltage* of IC3; make a note of that voltage, which can be anywhere from 0 volt to $\pm$ 10 millivolts. Remove the clip lead. Using your DMM, measure the exact value of R_C (4.02 ohm, 1%), minus your test lead resistance. Connect R_C to BP1 and BP2, push the TEST switch, S1, and adjust CAL potentiometer R4 so that the output from J1 (about 4.02 volts) indicates the exact value of R_C added to IC3's offset voltage. Re-assemble the cabinet and the milli-ohm adapter is complete and ready for use.

In actual use, connect the unknown resistance (R_x) at the exact points on its leads from which you want to measure, take the reading, and *subtract* IC3's offset voltage for the proper value at 1 ohm/volt. That procedure will ensure the best possible accuracy. Ω

FOR AS LONG AS SHORTWAVE LISTENING has been popular, it has been difficult to match a random-length antenna to the typical 50- or 75-ohm input impedance of most radio receivers. The impedance of a random-length shortwave antenna can vary from a few ohms to several hundred ohms, depending

on frequency. Matching that impedance to the input impedance of a receiver is possible, but it can be complicated.

Many different active-antenna designs provide effective matching, but many people believe that active antennas never work satisfactorily. An active antenna is typically a single-stage, wideband amplifier with a gain from 5 to 8 dB and a noise figure from 3 to 8 dB, making it a device of questionable merits. While active antennas provide some gain, they can also add noise to the signal, especially a signal at a low microvolt level.

TUNABLE SHORTWAVE ANTENNA

Improve your shortwave reception with this tunable antenna.

H.J. WECKE

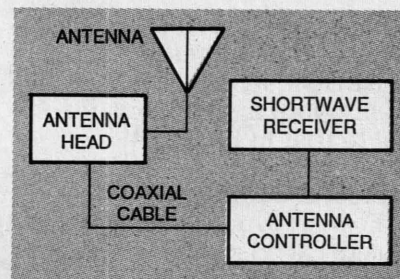
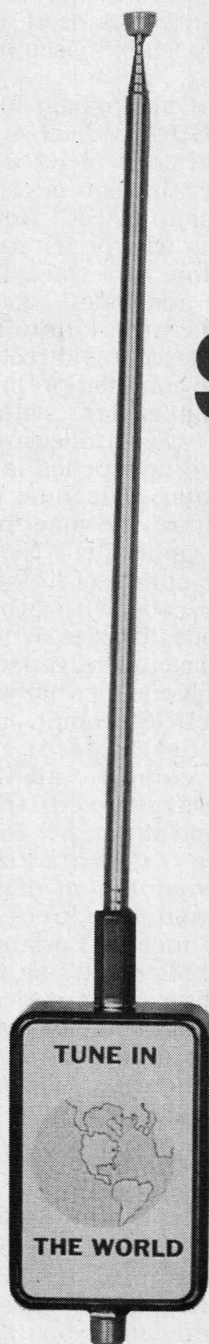


FIG. 1—THE TUNABLE SHORTWAVE antenna adds a tuned stage ahead of the 50- or 75-ohm input to your shortwave receiver.

Active antenna wideband amplifiers really lose out in the low microvolt region. The inherent noise usually kills a low-level incoming signal. Nearby transmitters might cause intermodulation in the active antennas as well. As a compromise solution, some older shortwave receivers had trimmer potentiometers on their front panels to provide impedance matches for the antenna at the received frequency. Sadly, modern receivers no longer have this control.

The author believed that a new method had to be designed to meet the following list of provisions:

- 1—No amplification should be needed, no noise should be generated, and no intermodulation should occur
- 2—It should have selectivity, but keep out strong adjacent signals
- 3—It should be small
- 4—It should be inexpensive, easy to build, and include only standard components

The solution was found by adding a tuned stage in front of the 50- or 75-ohm input to the shortwave receiver. A tuned stage has many advantages over the best wideband RF amplifiers, including preselection at resonance and elimination of both noise and intermodulation. The need for tracking the antenna from band to band is the only disadvantage this tuned stage has over a wideband active antenna. Several prototype antennas have been in service reliably for more than four years. They have been mounted on a balcony or a roof where they were exposed to a harsh climate. Figure 1 is a block diagram of the system.

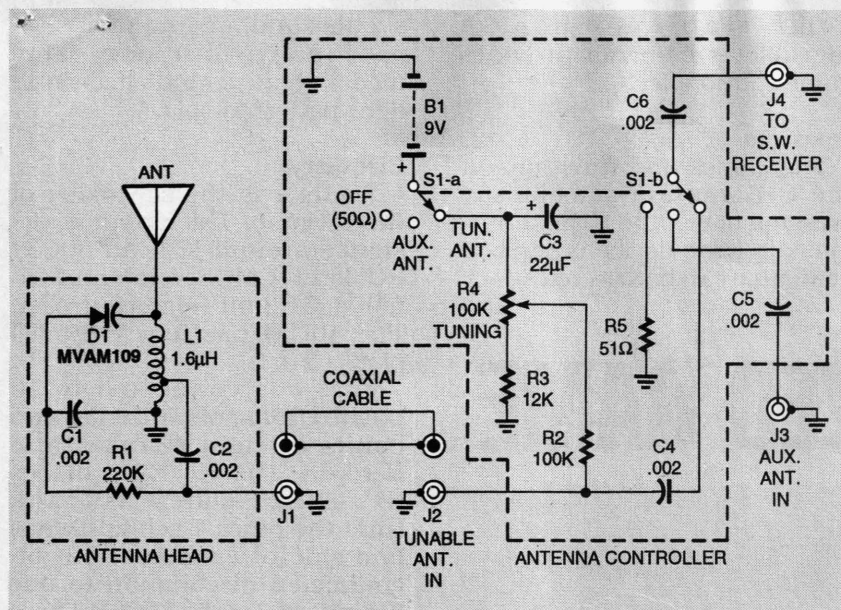


FIG. 2—TUNABLE ANTENNA SCHEMATIC. The variable element is Motorola MVAM-109 diode D1.

Circuitry

Figure 2 is the schematic for the tunable antenna system. A single monopole antenna about 3 feet long was chosen. The antenna, which has a measured capacitance of approximately 10 picofarads, becomes part of the antenna-head circuit that feeds the antenna controller. The variable element of the antenna head is a Motorola MVAM-109 diode (D1). The inductor value required to cover the shortwave spectrum from about 6 MHz to 18 MHz was calculated to be 1.6 microhenries. The inductor can easily be wound by hand on a 0.25-inch diameter form containing a ferrite slug.

The MVAM-109 requires a tuning voltage from 1 to 9 volts DC at a low current. That is provided by 100-kilohm linear potentiometer R4 located in the antenna controller section. Switch S1, when in the "off" position, places a 50-ohm termination on the tuned output. The other two positions of S1 select either an auxiliary antenna input or the tuned antenna input. A length of coaxial cable terminated with suitable connectors connects the antenna head, which might be located outdoors, to the antenna controller, which should be located in a convenient place near your shortwave receiver.

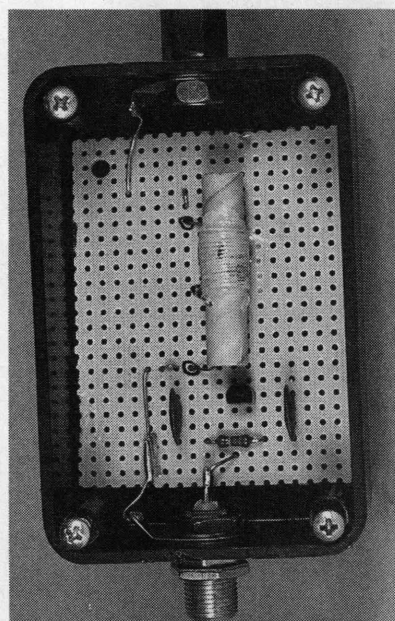


FIG. 3—THE ANTENNA SHOULD be mounted to the antenna-head case with a neoprene washer, and a bead of silicone sealant should be applied to the edge of the lid to keep out water.



FIG. 4—ANTENNA CONTROLLER unit. House the controller in a metal case to provide RFI shielding.

PARTS LIST

All resistors are 1/4-watt, 10%, unless otherwise noted

R1—220,000 ohms, 10%

R2—100,000 ohms, 10%

R3—12,000 ohms, 10%

R4—100,000 ohms, linear potentiometer

R5—51 ohms, 10%

Capacitors

C1, C2, C4—C6—0.002 μ F, ceramic

C3—22 μ F, 16 volts, electrolytic

Semiconductors

D1—Motorola MVAM-109 varactor diode (nominal capacitance of 460 pF at $V_R = 1.0$ V and $f = 1.0$ MHz) or equivalent

Other components

L1—1.6 μ H (wind by hand, see text)

J1, J2—panel-mount F connector

J3, J4—any coaxial connectors suited to your needs

B1—9-volt carbon-zinc battery (do not use alkaline)

S1—2-pole, 3-position rotary switch (optional, see text)

Miscellaneous: Antenna, two cases, silicone sealer, coaxial cable, No. 26 magnet wire, 1/4-inch form with ferrite slug, 9-volt battery snap connector

Note: The following items are available from Wecke Associates, P.O. Box 3822, Ottawa, Canada K1Y 4M5:

- A kit consisting of D1, L1, and 6 mm bolt, nut, and solder lug to accommodate a 30-inch automotive whip antenna—\$6.00 + \$2.00 S&H

- Assembled and tested antenna controller, antenna head, and aluminum mounting bracket with U-bolts and battery (does not include whip antenna or connecting cables)—\$65.00 + \$5.00 S&H

Payment required in US dollars. Canadian customers please add appropriate provincial taxes. Allow 3-4 weeks for delivery.

Construction

The antenna head can be hand wired on perforated construction board and installed in any weathertight case. Good, clean solder joints are important in the antenna head, which will probably be exposed to the weather. The layout of the parts and wiring is not critical—just avoid excessive lead lengths. The case selected for the prototype antenna head has an aluminum cover that can be used as a base for attaching a mounting bracket crafted from scrap aluminum. Coil L1 can be made by winding 16 turns of No. 26 magnet wire on a 1/4-inch diameter

Continued on page 90

TUNABLE ANTENNA

continued from page 61

ter form. Place a tap at the second turn from the grounded side of the coil, and insert a ferrite slug in the form. Coat the finished coil with clear nail polish or lacquer before soldering it into the circuit.

An F connector was installed at the bottom side of the antenna head for connecting the coaxial cable that runs between the head and the controller. The antenna itself should be mounted to the antenna-head case with a neoprene washer to keep water out of the unit. For the same reason, apply a bead of silicone sealant around the edge of the lid. The finished antenna head is shown in Fig. 3.

On the antenna-controller section, two-pole, three-position rotary switch S1 allows tuning, the use of an auxiliary antenna, or the ability to switch the unit off entirely. A two-position switch can be substituted if you do not want an auxiliary antenna. Also, because current consumption is so low, the switch can be omitted and the battery left in the circuit permanently where it might attain its rated shelf life. Use shielded cable to connect components S1, R4, and J2-J4 to the rest of the circuit.

Do not use an alkaline 9-volt battery because it could provide too much current for proper operation. Instead, install a carbon-zinc battery. Warning! Do not take the required 9-volt power from the internal circuitry of your shortwave receiver, even if the antenna controller is built into the receiver. House the controller in a metal case, because it will provide the best RFI shielding. Figure 4 shows the prototype controller.

When the unit is completed, allow a "burn-in" period for the tuning diode. Perform burn in by installing a battery, turning the unit on, and allowing the silicon crystal structure of the diode to settle with a few hours of operation. That step proved to be necessary for many of the units built. Ω

THE LED-HEAD

continued from page 48

to mount the circuit board in a suitable plastic box. The size of the box will, of course, be determined by your choice of power—battery or adapter. If you use the four cells, you can mount the holder in the bottom of the box; if you choose an adapter, mount the jack on the side of the box.

Where to put the Face

With appropriate packaging the Electronic Face can become:

- An novel desktop curiosity and attention getter.
 - A night light for children's rooms.
 - A holiday ornament to be hung on the walls, doors or even a Christmas tree.
 - An novel automotive accessory to keep the guy in the car behind you amused when you are stopped at a red light.
- You'll probably think of many more possibilities. Ω

RELAY OUTPUT CIRCUITS

continued from page 54

ever, the period is determined by the switch position to either electrolytic capacitors C1 (minutes) or C2 (seconds), until the relay turns off.

At that time, the control contacts reopen and break the power connections to the circuit. The timing cycle is then complete. This circuit can be turned off part way through its timing cycle by pushing RESET switch S3.

Standard aluminum electrolytic capacitors have very wide capacitance tolerance values (typically -50% to +100%). Moreover, they exhibit relatively large and unpredictable leakage currents. Consequently, their use in simple circuits such as those shown in Figs. 13 and 14 make them unsatisfactory for precise timing of relay contact functions. Also, they are unable to time periods longer than about 15 minutes.

Figures 15 and 17 show two accurate, long-period relay timer-control circuits whose functions do not depend on electrolytic capacitors. Film dielectric capacitors have been substituted. In both of those circuits, IC1 is configured as a free-running astable multivibrator.

In Fig. 15 schematic, a two-range, 1 to 10 minute and 10 to 100 minute relay timing control circuit, the astable frequency is divided down by IC2, a CD4020B CMOS, 14-stage, ripple-carry binary divider. Consequently relay RY1 turns on as soon as switch S1 is closed, and it turns off again when the 8192nd astable pulse arrives. This provides time periods from 1 to 100 minutes, depending on the position of switch S2.

Figure 16 shows the functional and pinout diagrams of the CD4020B 14-stage binary counter. All counter stages are master-slave flip-flops. The state of a counter advances one count on the negative transition of each input pulse, and a high level on the RESET line resets the counter to its all zeros state. All inputs and outputs are buffered.

The Fig. 17 circuit is similar to that of Fig. 15 except that an additional decade-divider stage is substituted for position 3 of switch S2. This gives a maximum division ratio of 81,920, making it possible to time for periods of up to 20 hours. This circuit is found in battery chargers and area security lighting systems with time-controlled turn-off.

Figure 18 shows the functional and pinout diagrams of the CD4017B decade counter IC, a five-stage Johnson counter with 10 decoded outputs. Inputs include a CLOCK, a RESET, and a CLOCK INHIBIT signal.

Time-delay relays are coil and contact EM relays with built-in timing circuits that delay contact openings or closings for a preset interval. The time delays are generated by an internal RC circuit or by counting cycles on a 50- or 60-Hz AC power line. The time delays are set with either a knob or a DIP switch on the relay. Ω

**BRANCO JUSTIC and
PETER PHILLIPS**

NIGHT-VISION SCOPES WERE DEVELOPED as military surveillance devices to permit viewing enemy activities and aiming weapons at night without revealing the observer's presence. The sensitivities of their principal components, image tubes, have been improved with fiber-optic lenses, more gain stages and better photocathodes. In addition, miniature, solid-state, extra high-voltage power supplies have reduced their size, weight and power needs.

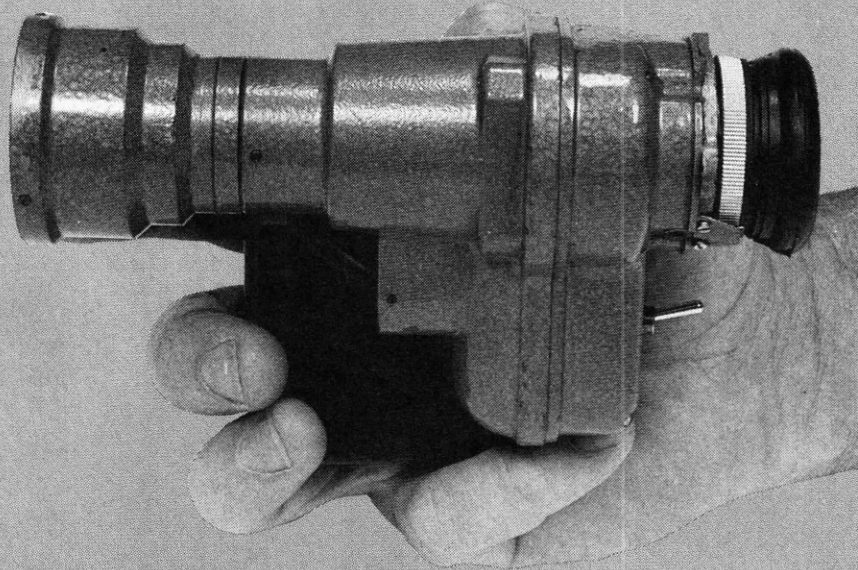
Two night-vision scopes are described in this article. One is passive, meaning that it will work in faint natural light, and the other is active, meaning that it requires supplemental infrared illumination. They include surplus first-generation imaging tubes. Although they have been superseded by more advanced devices, they will, nevertheless, provide adequate sensitivity for most hobbyists and science experimenters.

The active scope will permit police to observe suspected criminal activity at night and citizens to monitor their homes or property without being detected. The scope will also permit hunting, nature study, marine navigation, and many other nighttime applications. The active scope is suitable for some of these activities, but the scene must be illuminated by an infrared source. Neither will disturb the eyes' adaptation to darkness.

The first night vision scopes were designed for use by forward observers, snipers, aviators, and tank crews. Some that were made as monoscopes to mount on rifles looked like the devices shown in Fig. 1; others were made as binoculars. The most sensitive passive units are called *starlight scopes*. *Night-vision goggles* are lightweight binoculars for helicopter crews that mount on their helmets.

Active night-vision scopes, such as the one shown in Fig. 2, depend on infrared illumination from sources such as lasers for aiming artillery, guided mis-

NIGHT-VISION SCOPES



***View a scene in near total darkness
with a passive night-vision scope,
or illuminate it with infrared
for an active scope***

siles, and "smart" bombs. Night-vision systems were considered military secrets for many years. After they were declassified, they could be sold as military surplus and commercial versions based on the technology were offered for police surveillance and as nighttime marine navigational aids at prices that often exceed \$2000.

Both of the night-vision scopes described in this article are based on military surplus equipment that includes both an image tube and optics. The parts for the active unit cost \$90, and parts for the active unit cost \$220.

Night-vision scopes

Figure 3 illustrates a typical night-vision scope. The objec-

tive lens, positioned at the cathode end of the tube, focuses the image on the photocathode. It is selected for its intended application—long-distance or short-range viewing. The eyepiece at the anode is for viewing the enhanced image. It is a simple lens that magnifies the image on the screen. It can be removed and replaced by a television camera, camcorder, or film camera for transmitting or recording the image.

The image tubes are the hearts of the night-vision scopes. Before you start building one (or both), you might want to learn more about how they work. See the sidebar entitled "Image Converter and Intensifier Tubes."

The only electronics needed

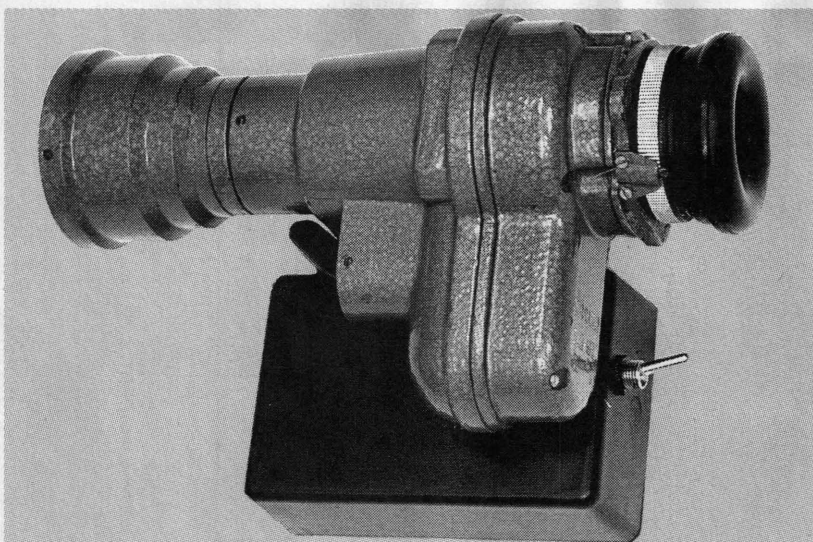


FIG. 1—A PASSIVE NIGHT-VISION MONOSCOPE made from half of a Russian night-vision binocular with an image intensifier tube and all optics.

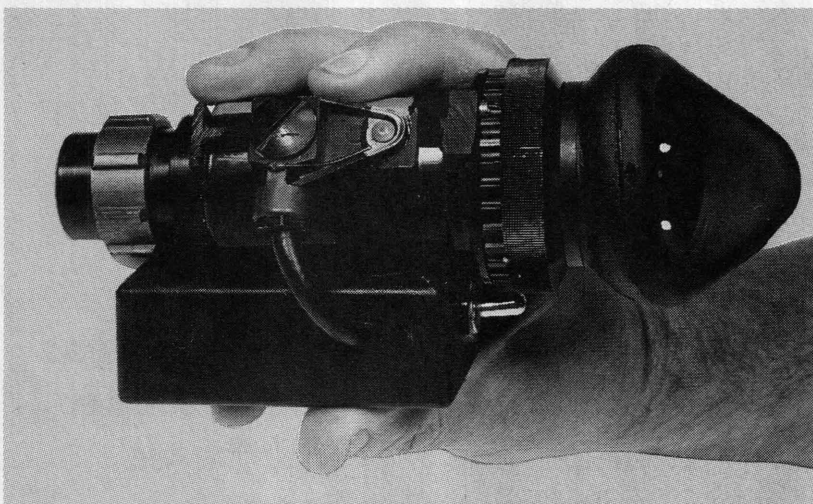


FIG. 2—THIS ACTIVE NIGHT-VISION SCOPE requires an infrared illumination source but it works from the same power supply as the scope in Fig. 1.

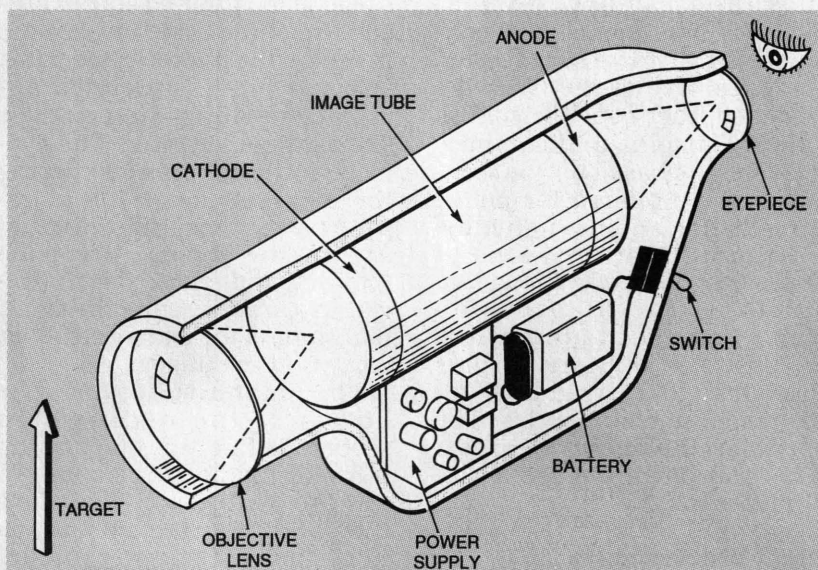


FIG. 3—ALL NIGHT-VISION SCOPES have objective and eyepiece lenses, an image tube, a high-voltage power supply, and a battery.

in both projects described in this article is a high-voltage power supply capable of providing a typical working voltage of 13.5 kilovolts. This efficient and compact regulated supply operates satisfactorily from a 9-volt, alkaline battery. The current drain of both tubes described here is small, so that their power consumption is low.

The compact power supply is built into a small plastic project case that is fastened directly to the surplus night-vision scope that contains the imaging tube. The Russian-made monocular viewer shown in Fig. 1 is actually one half of a binocular. It is complete with an objective lens and an eyepiece. This assembly includes a first-generation, single-stage *image intensifier tube*.

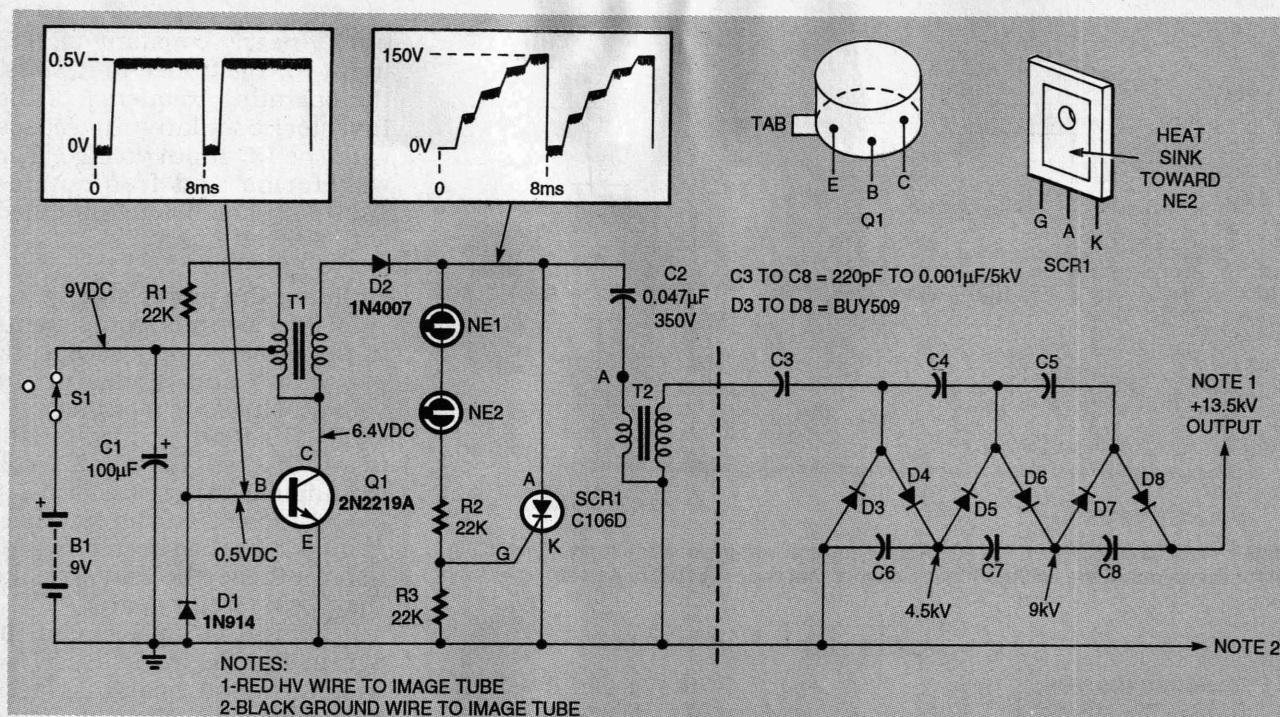
The active night-vision scope shown in Fig. 2 contains a single-stage *image converter tube*. Instructions on how to make several different low-cost infrared illumination sources are described in this article.

Active military night-vision weapons aiming systems typically include an infrared-emitting laser. It pinpoints the target for a heat-seeking weapon or for aiming other kinds of guns or missiles while also acting as a non-visible searchlight for the observer (bombardier or gunner) with an active scope. Various systems have been built for use on land, in the air, or on the sea at night.

Infrared-sensing missiles and "smart" bombs actually "home" on the IR-illuminated target which has been identified by the observer who directs the laser beam and watches it with the active scope. Needless to say, aiming and firing must be fast because enemy gunners with active scopes can also see the laser illumination and take evasive action or retaliate.

Power supply design

Figure 4 is the schematic for a high-voltage power supply that will power both night-vision scopes described here. It produces about 13.5 kilovolts from a 9-volt battery. The tubes draw about 20 milliamperes so about



36 hours of useful life can be obtained from a 9-volt alkaline battery. The output voltage will remain essentially constant for battery voltages of 6 to 12 volts.

The power supply has three sections: the inverter, the converter, and the voltage multiplier. The inverter section, a ringing-choke oscillator, consists of transformer T1, resistor R1, diode D1, and transistor Q1. Resistor R1 provides bias current for starting the oscillator, and it also supplies the feedback to maintain oscillation.

Diode D1 protects the base-emitter junction of Q1 when the base voltage swings negative. The oscillator operates at about 120 Hz, set principally by the transformer. The resulting AC voltage at the primary of T1 is stepped up by the secondary turns. The secondary voltage, which is rectified by diode D2, charges C2 through the primary (low-resistance) winding of transformer T2.

When the voltage across C2 exceeds the breakdown voltage of the two series-connected neon lamps NE1 and NE2 (about 150 volts) the lamps turn on. This conduction triggers SCR1, and C2 is quickly discharged through SCR1 and the

LIGHT UNITS

The standard (SI) unit of light intensity is the lux, and a typical recommended minimum illumination level for a video camera is around 50 lux. However, some video cameras will work with light levels as low as 1 lux. With this frame of reference, you can get a better idea about how sensitive night-vision systems are.

The recommended light level for the safe operation of these tubes is from 50 millilux down to 500 micro-lux (0.0005 lux)—very low light levels. Table 1 lists familiar light levels and their light intensity in lux units. By contrast, a typical first-generation image intensifier tube will not be damaged if the illumination level is as high as 100 millilux (one tenth of a lux), but tube life will be reduced if it is exposed for long periods to that light level.

Sensitivity is the most important specification for a night-vision scope. An average scope has a sensitivity of 10 millilux (mlx), a value that makes it effective in night light-

primary winding of T2. When C2 is discharged, the lamps extinguish, SCR1 turns off, and the charge cycle starts again.

During the discharge cycle of C2, a pulse with a peak-to-peak voltage of 4.5 kilovolts is produced at the secondary of trigger transformer T2. This pulse

TABLE 1
RELATIVE VALUES OF
NATURAL LIGHT SOURCES

Source	Illumination (lux)
Direct sunlight	100,000
Bright sunlight	10,000
Overcast day	1,000
Very dull day	100
Twilight	10
Deep twilight	1
Full moon	0.1
Quarter moon	0.01
Starlight	0.001
Overcast starlight	0.0001

ing from a quarter moon. By contrast, a very sensitive night-vision scope can operate at 1 mlx, which corresponds to starlight conditions. For this reason, very sensitive scopes are called "starlight" scopes.

is applied to the three-stage Cockcroft-Walton or Greinacher voltage-multiplier circuit consisting of diodes D3 to D8 and capacitors C3 to C8.

The multiplier triples the 4.5-kilovolt input to provide the 13.5-kilovolt output with very low current. The capacitors and

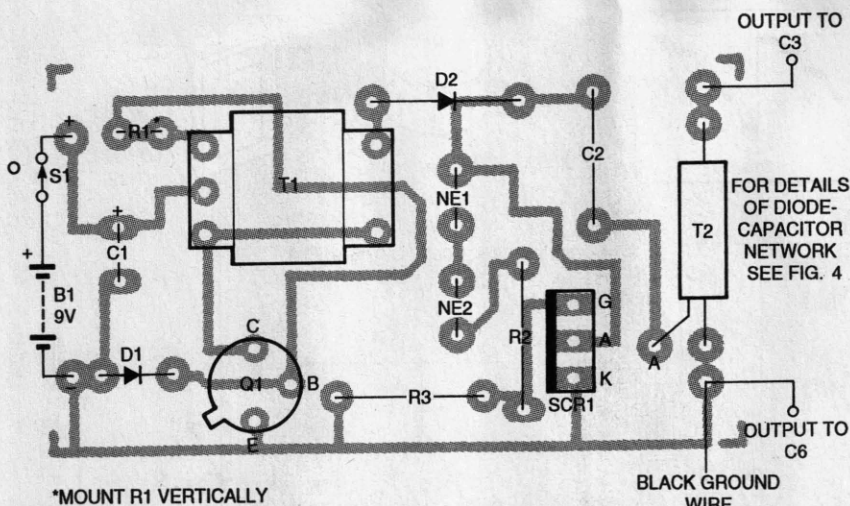


FIG 5—PARTS PLACEMENT DIAGRAM for the high-voltage power supply. Refer to Fig. 4 for the layout of the voltage-tripler network that is connected to its output.

PARTS LIST

Resistors are ¼-watt, 10%

R1, R2, R3—22 kilohms

Capacitors

C1—100 μ F, 25 volts, aluminum electrolytic

C2—0.47 μ F, 350 volts, polyester

C3, C4, C5, C6, C7, C8—220 pF, 5 kV breakdown, ceramic

Semiconductors

D1—1N914 silicon signal diode

D2—1N4007, 1000V, 1A silicon diode (DO-41 case), Motorola or equiv.

BY509—silicon diode, 10 kV, 3 mA, Philips or equiv. (rating must be greater than 6 kV, 2 mA)

2N2219A—NPN transistor, Motorola or equiv.

MCR106-6 (C106D)—SCR, 400 V, 4 A in a T-126 package, Motorola or equiv.

Other components

NE1, NE2—neon tubes, miniature

T1—transformer, inverter, 3 kilohm CT, iron core, audio, miniature PC mount

T2—transformer, trigger type, 250 V primary, 6 kV secondary

Miscellaneous printed circuit board; passive night-vision monoscope with imaging tube and optics (see text); active night-vision scope with converter tube (see text); plastic project box (see text); 9-volt alkaline transistor battery, 9-volt transistor battery clip with wires; miniature toggle

switch, PCB mounted; plastic sleeving; RTV silicone potting compound; tinned copper wire, 22 AWG; insulated hookup wire, 22 AWG, solder.

Note: The following items are available from Oatley Electronics, P.O. Box 89, Oatley, Sydney, NSW, Australia 2223: Phone 011 61 2 579 4985 (Time zones USA—East 7 PM to 2 AM, Central 6 PM to 1 AM, Mountain 5 PM to 12 AM, Pacific 4 PM to 11 PM), Fax 011 61 2 570 7910. Mastercard and Visa accepted with telephone or fax orders, international bank drafts and money orders accepted by mail. Customers please include phone and/or fax number.

• Complete kit of passive night-vision scope and parts for the HV power supply—\$220.00

• Complete kit of active night-vision scope and parts for HV power supply—\$90.00

• Kit of parts for HV power supply only—\$24.00

• Kit of non-standard parts (T1, T2, NE1, NE2, and C2)—\$8.00

Include \$15.00 for shipping and handling, \$6.00 for air mail from Australia.

put is nearly constant for DC input voltage from 6 to 12 volts, the operating frequency of the inverter/oscillator increases with the DC input voltage. The waveforms and frequencies shown on Fig. 4 were obtained with a 9-volt input.

Building the power supply

All the electronic components of the power supply except the capacitors and diodes in the voltage tripler are mounted on a $1\frac{15}{16} \times 1\frac{1}{15}$ -inch printed-circuit board that will fit with a 9-volt rectangular battery inside a $2 \times 3\frac{1}{4} \times 1$ -inch plastic project case. A foil pattern has been provided here for those who want to make their own circuit boards.

Refer to parts placement diagram Fig. 5. Insert SCR1 so that its metal heatsink faces neon lamp NE2. When inserting electrolytic capacitor C1 and diodes D1 and D2, observe their polarities. Mount resistor R1 vertically on the circuit board. Solder all components and trim excess lead lengths.

Refer back to the schematic Fig. 4, and wire the leads of capacitors C3 to C8 and diodes D3 to D8 together mechanically to form a rigid unit according to the schematic. Keep all exposed lead lengths about $\frac{1}{4}$ -inch long. Then solder the network together as rapidly as possible to avoid applying damaging excessive heat to the capacitors and diodes. The cathodes of some diodes are identified with a red dot on the cathode lead.

Figure 6 shows the completed power supply with the tripler network to the right of the board. Battery B1 and switch S1, both off-board components, are not shown. Solder one lead from C3 and one lead from C6 in the tripler network to the terminal points on the circuit board, as shown in Fig. 5. (The tripler will be potted in silicone after the system has been tested.)

NOTE: The image converter tube in the active night-vision scope shown in Fig. 2 requires a positive ground. If you build this unit, reverse diodes D3 to D8 to convert the supply from one with a positive to a negative output with respect to ground.

diodes must be rated to withstand at least 4.5 kilovolts. For this reason, the capacitors are all rated at 5 kilovolts, and the diodes are high-voltage units.

The neon lamps regulate the output so that the voltage applied to the primary of T2 is constant at 150 volts, peak. Although the power supply out-

Mechanical assembly.

Drill a hole in the body of the plastic project case for mounting the miniature toggle switch S1. The location of the switch is not critical, but it should not interfere with the other components. In the passive night-vision scope, it was positioned at the end of the case facing the eyepiece.

Drill the hole in the case for mounting it to the passive scope body with a single screw. The hole for this screw, already drilled and tapped in the body of the scope, is located under the coupling bracket. Drill another hole large enough to pass the power cable to the image tube. Its location will depend on the project you build. Mount switch S1 in the sidewall of the case. Fasten the case to the passive viewer with a screw. If you build the active viewer, cement the case to the scope body with epoxy, as shown in Fig. 2.

Cut the supply lead from the scope about 6 inches long, and strip back about 1 inch of the jacket to expose the braid and the insulated central conductor. Twist the braid into a lead and insulate it with a length of plastic tubing. Insert the cable lead into the project box, solder the braid to the ground connection of the circuit board, and solder the inner conductor to the high-voltage terminal of the tripler network, as shown in Fig. 4.

Verify that there are no short circuits in the construction of the tripler and that the leads are spaced by at least $\frac{1}{4}$ -inch from each other. Wire toggle switch S1 in series with the positive lead from the battery clip. Cut the battery leads from the circuit board so they are long

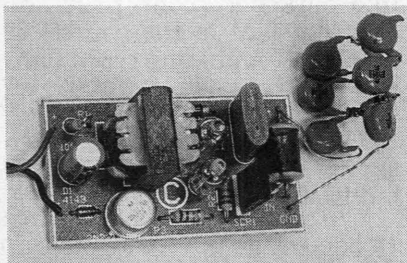


FIG. 6—COMPLETE HIGH-VOLTAGE power supply circuit board with the voltage-tripler, diode-capacitor network shown unpotted at right.

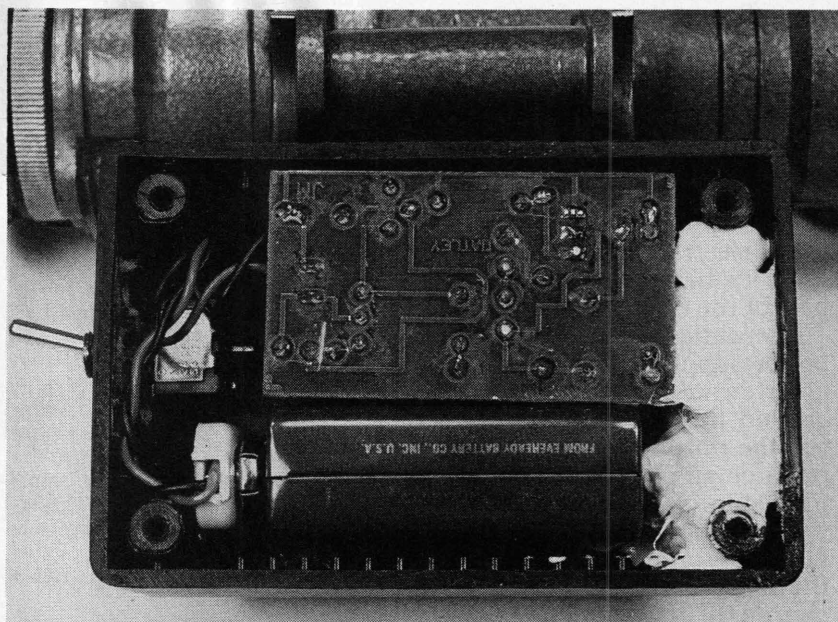
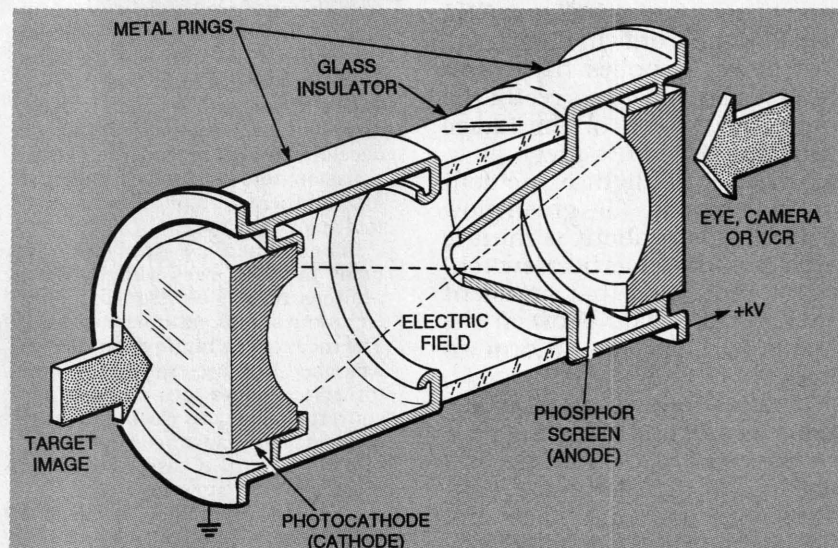


FIG. 7—POWER SUPPLY CIRCUIT, battery, and switch in the project case. The tripler network potted in silicone is shown as the white patch at right.



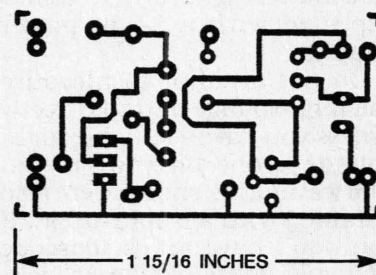
CUTAWAY VIEW OF SINGLE-STAGE image tube showing the position of its key parts.

enough to permit lifting the circuit board out of the case. Insert the battery and wiring in the case, but leave the tripler network and circuit board outside temporarily.

Test and checkout

After rechecking your work and verifying the polarities and orientation of all components, snap the battery to the battery clip. The current drain on a 9-volt alkaline transistor battery should be about 20 milliamperes in a correct circuit.

WARNING! The power supply described in this article produces an output voltage of



FOIL PATTERN FOR POWER SUPPLY circuit board.

about 13.5 kilovolts, and is capable of giving you a startling electric shock. While not normally life-threatening, it can have a temporarily debilitating effect. Consequently, treat it

with respect and always make sure switch S1 is off and the capacitors are discharged before handling the circuit board.

When you switch on S1, a corona discharge might appear around the tripler network. This is non-destructive, but avoid electrical shock by keeping your hands away from that part of the circuit.

Alternatively, you can test the power supply by placing a wire connected to the circuit's ground bus close to the high-voltage output lead. It should produce an arc as much as 1/4-inch long.

When the switch is off, connect the ground wire directly to the high-voltage output to discharge all the capacitors. Because of the short-duty cycle discharge pulses, the illumination of neon lamps NE1 and NE2 will be visible only in darkness.

If power supplies have been built and installed correctly, the phosphor screen of the image tube should emit a green glow, whether or not light is incident on the cathode. The green glow persists for a about a minute after the power has been switched off, indicating that sufficient voltage still exists between the anode and cathode to form an image.

If, after following the directions closely and checking your workmanship, the scope still doesn't work, check the voltages at the base and collector of transistor Q1 with a digital voltmeter. The values shown on the schematic, Fig. 4, are DC values expected with a 9-volt power supply.

Do not attempt to measure the high-voltage output directly unless you have a suitable high-voltage probe on your meter. The waveforms shown were also obtained with a 9-volt supply. If you don't have an oscilloscope available, measure the AC voltage at the test points shown on Fig. 4.

The reading on most digital voltmeters will be RMS values, but they will give you a valid indication if there is a signal. The AC voltage at the cathode of D2 (to ground) of the prototype measured about 45 volts RMS

IMAGE CONVERTER AND INTENSIFIER TUBES

An image tube is a vacuum tube that converts an image in one part of the electromagnetic spectrum to another part of the spectrum, usually with an increase in intensity. If, for example, you want to convert a near-infrared (IR) image to a visible light image, you would use an *image converter tube*. This is the tube in the active night-vision scope described in this article. However, the scene must be illuminated by an infrared source if the scope is to function.

However, if you just want to intensify a scene in darkness you would use an *image intensifier tube*, the kind included in the passive night-vision scope described here. The figure shows a simplified cutaway view of this kind of tube. While it is also sensitive to near IR, it is still called an image intensifier.

Both tubes have photoemissive input surfaces at their cathode ends that form an electron image of the scene being viewed when a voltage of 4 to 20 kilovolts is applied between the cathode and anode. Photons from the scene or target are focused on the photocathode which triggers an avalanche of electrons. The anode accelerates the cascaded electron emission from the photocathode and focuses them on a phosphor screen to display the image.

Image tubes are now made in many different sizes with different photocathodes, electron focusing techniques, and gain mechanisms. The tubes in the night-vision scopes described here are examples of first (or zero) generation, single-stage converters that are electrostatically focused. More advanced tubes obtain higher gain by cascading two or more converter stages.

For night-vision applications, the earlier generations of image tubes had simple glass lenses and S25 photocathodes that closely match the peak sensitivity of the human eye. These tubes produce a high-definition black and green image. The latest versions now have fiber-optic lenses, gallium-arsenide photocathodes, and gain stages called *microchannel plates*.

In night-vision scopes, the image must be formed on the cathode by an objective lens similar to a camera lens, and the scene is viewed at the anode end with an eyepiece.

It is important to remember that no passive image intensifier tube will work in complete darkness. It must gain enough light from natural background sources to work. These might only be barely perceptible even to the night-adapted human eye. The sources could be faint sky glow on a cloudy night, starlight on a clear night or moonlight.

on a DMM. The AC voltage at the base of Q1 was about 0.45 volt RMS.

When the scope is working properly, switch off the power and wait for the tube to discharge completely. Insert the tripler section carefully inside the case as shown at the right side of Fig. 7. Encapsulate it with neutral-cure, room-temperature vulcanizing (RTV) silicone potting compound to prevent high-voltage corona and discharge, which increases with relative humidity. The compound will also fasten the tripler network inside of the case.

Viewing with the scope

The lens of the surplus Russian passive night-vision scope specified in this article is focused to infinity, making it useful for viewing images more than a few meters away. By loosening a small locking screw, the lens can be adjusted. This will permit viewing objects more than 100 meters away under near starlight illumination.

Imaging tubes will be damaged if they are exposed to bright light for long periods. Don't use either scope in sunlight, or even in well lighted rooms. Always cover both ends of a night-vision scope with suitable lens caps when it is not in use to keep the imaging tube in darkness. The monocular passive scope offered by the source given in the parts list has a rubber lens cap that can be snapped in place. It also has a focusing eyepiece.

Both of the night-vision scopes will detect IR energy, so they can verify the operation of stereo and TV remote controls. In a darkened room, point the emitting face of the control at the scope. A pulsing green light will be seen when any of the remote control's keys are pressed. A TV remote control can also serve as a temporary IR illuminator.

IR light source

A better IR source can be made by covering a flashlight with an IR filter. You can pur-

Continued on page 73

RADIO AMATEURS HAVE BEEN communicating with two-way television for years, but their equipment was so expensive that it cost more than most amateurs could afford. However, recent advances in solid-state technology have drastically reduced the prices of the necessary components for amateur television (ATV) and made them available to a far wider group of amateur operators.

Today, many key components in ATV systems that were once either prohibitive in cost or unavailable are now stock items. These include VCRs, camcorders, low-priced, quality color TV receivers, and solid-state video cameras. Credit-card-sized cameras based on charge-coupled devices (CCDs) are now available for less than \$200.

All of those components, together with the transceiver and transmitters described in this article, make the mobile, battery-operated TV station a reality. You can package a complete station in a case weighing less than 5 pounds that can easily be operated in the field.

In addition to two-way communications, the ATV projects presented in this article can be applied to radio-controlled model aircraft, boats, and cars. A transmitter and a small CCD camera placed in the vehicle act as "eyes" for the radio-controlled model; the view from the platform can be seen on a TV receiver. You can also build wireless video cameras, surveillance devices, and carry out unobtrusive nature studies. All this is possible with a package that fits in the palm of your hand and operates from AA power cells.

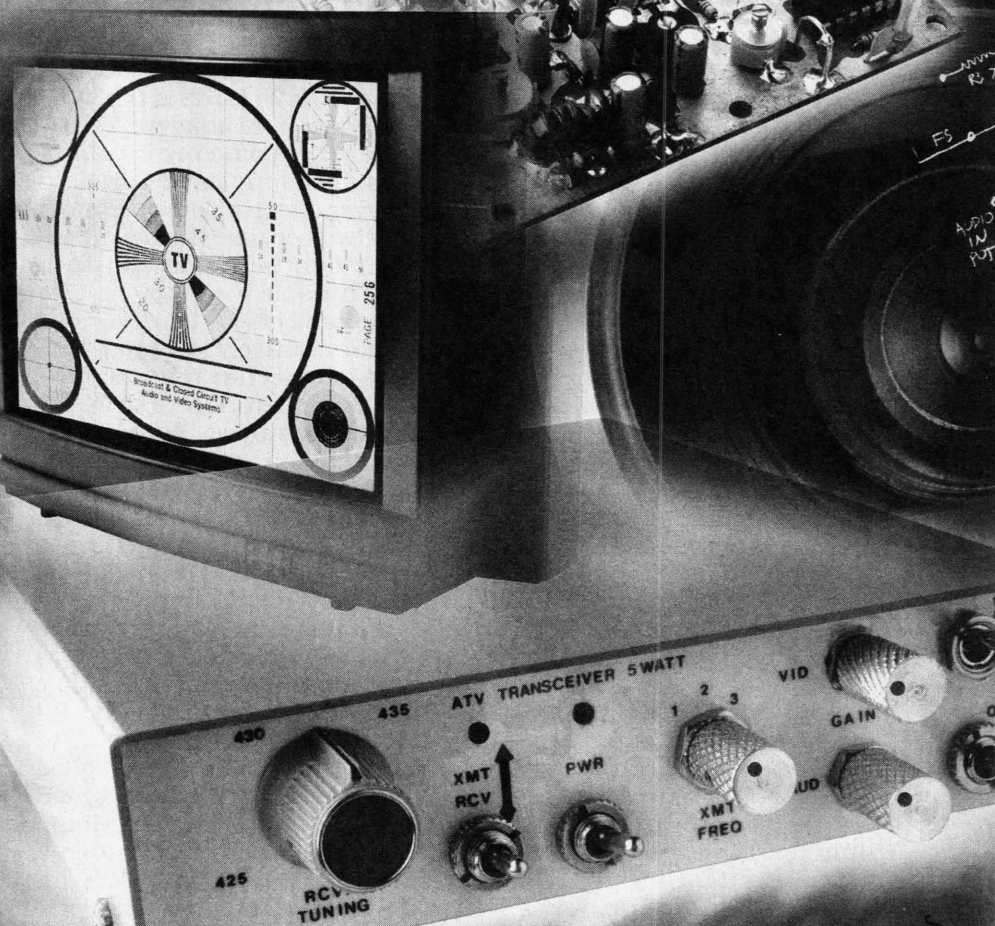
This is a two-part article; the first part describes the operation of the projects and the second part will cover the construction of them. There are four related ATV projects:

- Complete ATV transceiver with a 5-watt, three-channel transmitter, downconverter, line sampler, and RF switching relay
- 5-watt, three-channel transmitter only

Build this Amateur TV Station

Build your own 2-way TV station that permits base and mobile operation.

WILLIAM SHEETS AND RUDOLPH GDA



- 3/4-watt, single-channel Mini ATV transmitter
- 3/4-watt, single-channel ATV Jr. transmitter (it is the same as Mini ATV but without audio capability)

The first project is a complete two-way amateur TV station, with a three-channel transmitter, an integral receiving con-

verter, antenna relay, and test circuitry, all contained on a $3\frac{1}{2} \times 4$ -inch PC board. An integral heatsink that cools the RF power amplifier and modulator adds another $\frac{1}{2}$ -inch, making the overall size $4 \times 4 \times 1$ inch.

Because some readers might want to build only the transmitter section, the PC board is de-

signed so that the receiving section (the downconverter), relay, and line sampler (the test circuitry) can be omitted. That will result in a three-channel, 5-watt ATV transmitter measuring only $2\frac{1}{2} \times 4$ inches, including the heatsink.

The transmitter can be made even smaller if transmitter power less than 5 watts is acceptable. The low-power transmitter, which measures $1\frac{3}{4} \times 2\frac{3}{4}$ inches, generates a 0.5 to 1-watt output, and is powered from an 8- to 14-volt source. Because it can be powered by a 9-volt battery, this smaller version is suitable for many applications in radio-controlled model aircraft, boats, cars, and robots. Its circuitry is similar to the 5-watt version except that it has no high-power RF amplifier and modulator. Table 1 lists suggested applications for the different ATV units.

A valid Technician Class amateur license (which does not require a knowledge of Morse code) is required to operate these transmitters legally in the United States. Also, you should know that it is illegal to transmit on commercial frequencies with these units. If you do not have a license, you can operate these transmitters only into nonradiating dummy loads for test purposes. However, you can legally construct these units and you can listen to the transmitter, but you will need a license to transmit. Outside the U.S., similar laws might apply—check the regulations of the country you are visiting. A no-code amateur license is suitable, and if you can construct these projects successfully, you should be able to pass the required exam for an amateur license easily. Call your local amateur radio club, or a licensed radio amateur for details. You can also write to the American Radio Relay League (ARRL), Newington, CT 06111 for details. (203) 666-1541.

ATV transmissions can be received on a standard TV receiver with a simple downconverter that converts signals in the 420 to 450 MHz band down to a suitable IF frequency. In the lower

TABLE 1—ATV APPLICATIONS

5-Watt Transceiver	ATV base station ATV portable and mobile operations ATV video handie-talkies
5-Watt Transmitter	ATV base with separate down converters
3/4-Watt Transmitter	R/C video links Surveillance applications Remote sensing Wireless cameras

VHF range, Channel 3 (60 to 66 MHz) or Channel 4 (66 to 72 MHz) are the IF frequencies most commonly used. A TV tuner usually can be modified to tune down to 420 MHz because all UHF TV tuners will tune as low as Channel 14 (470 MHz). Some cable-ready sets can also tune to certain UHF amateur frequencies. Channel 60 (cable), at 439.25 MHz, is the most commonly used amateur TV frequency. A downconverter provides the simplest approach, however. It can also be the most effective, because the downconverter can be made continuously tunable.

The projects to be described include relatively low-priced RF transistors characterized for 450-MHz operation to deliver 1 or 5 watts output. With a suitable antenna, a 30- to 40-mile range can be obtained, depending on factors such as height, terrain, local noise and interference levels, and path characteristics. A small antenna, such as simple 6-inch quarter-wave whip, will reduce the range. In general, simple antennas give unspectacular results. An efficient antenna at 450 MHz doesn't have to be large; a good Yagi antenna with over 10 dB gain need only be a few feet long.

The transceiver includes a three-channel 5-watt crystal-controlled transmitter that produces both picture and sound, a low-noise, tunable downconverter covering the 420 to 440 MHz ATV frequencies, a sampling circuit to monitor or test the transmitter output, and a changeover relay to perform antenna (RF) switching. The circuit draws less than 50 milliamperes when receiving and 800 to 900 milliamperes when transmit-

ting. Typically, the transmitter produces 6 watts peak output into a 50-ohms load.

Circuitry

Figure 1 is the ATV transceiver block diagram. Notice that the transmitter section is enclosed in dashed lines. The crystal oscillator produces a signal of about 110 MHz. The exact frequency depends on the crystals installed. Any three crystals whose frequencies are within 2.5 MHz of each other are suitable. They will allow a 20-MHz frequency spread between the three output frequencies. A PIN-diode network (Q1, D1, D2, D3) selects the appropriate crystal.

A tuned network couples the oscillator's second harmonic to doubler Q2, which produces a 220-MHz output. That is coupled to a second doubler, Q3, which again doubles the frequency to 440 MHz and produces 50 to 75 milliwatts of RF energy. A double-tuned network couples this energy to Q8 which can produce nearly 1 watt of RF energy. In the $\frac{3}{4}$ -watt transmitter, Q8 is the final RF amplifier. (It is labeled Q4 in that version.)

Amplifier Q8 drives Q9 to produce an output of about 6 watts during modulation peaks. Video modulator Q10 is in series with the collector voltage supply to Q8 and Q9. It modulates the V_{CC} supply to Q8 and Q9 with the video signal it receives from amplifiers Q6 and Q7. A small sample of the RF output voltage is rectified by D11 and fed to emitter-follower Q14. The output of Q14 is the modulation envelope of the transmitted signal. The envelope can be observed with an oscilloscope, or it can drive a monitor so that you can check picture quality.

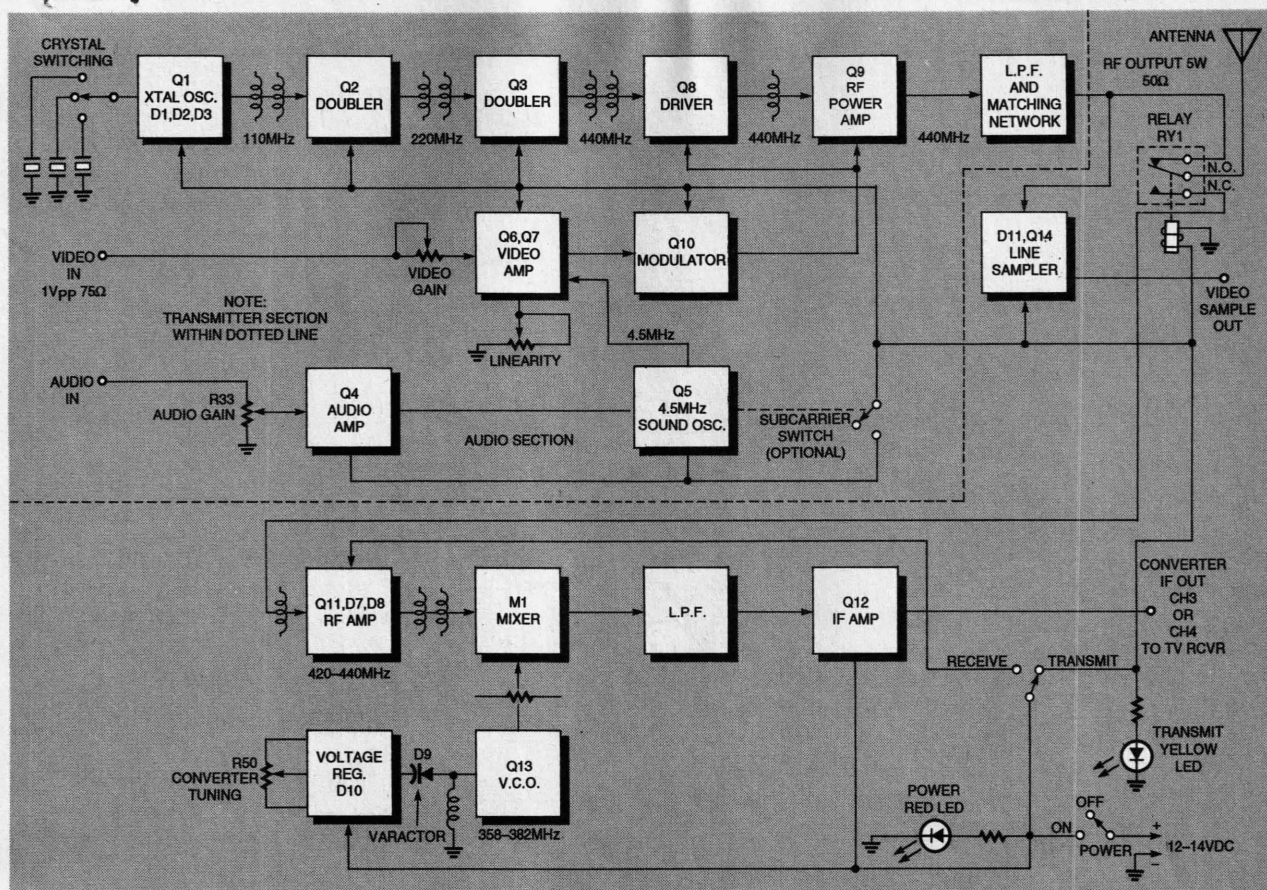


FIG. 1—ATV TRANSCEIVER BLOCK DIAGRAM. The transmitter section is enclosed in dashed lines.

The transmitter output is fed to relay RY1 which is activated by the transmitter V_{CC} line. The relay is used solely for RF switching. In the receive mode, the relay is de-energized and RF from the antenna is fed to down-converter RF amplifier Q11. The RF stage has a noise figure of less than 1 dB, a single-tuned input, and double-tuned output. Gain is about 20 to 23 dB. Signals from the RF amplifier are fed to double-balanced mixer M1 where they are mixed with the local oscillator (LO) signal from Q13. Voltage-controlled oscillator (VCO) Q13 has about a 15-MHz tuning range.

Varactor diode D9 is fed a bias voltage from tuning potentiometer R50. During transmit, DC voltage is removed from the RF stage. That reduces down-converter gain by about 50 dB. It also allows the downconverter to tune to the transmitted signal, which is useful for monitoring. The IF output of mixer M1 (60 to 72 MHz) feeds a lowpass filter that rejects UHF signals

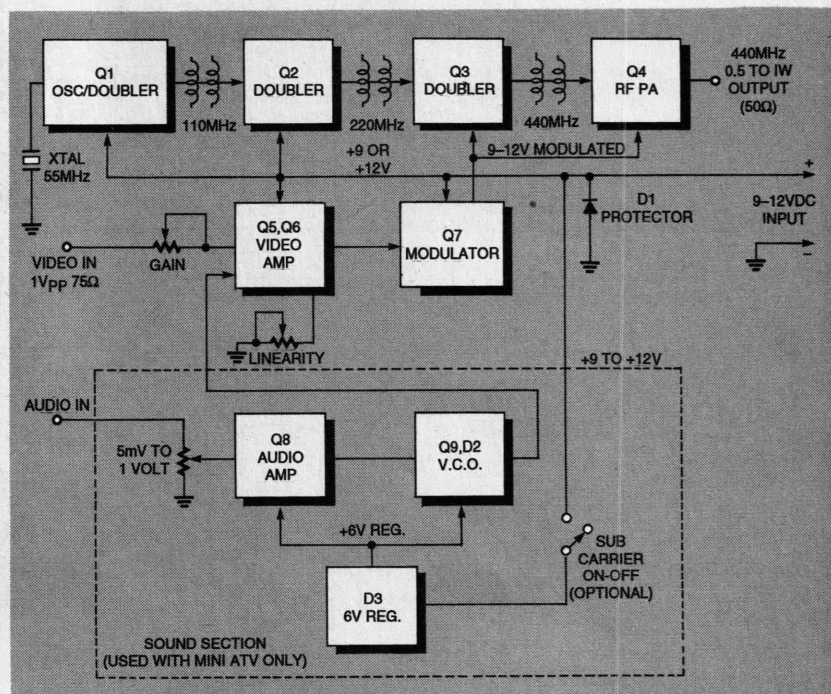


FIG. 2—BLOCK DIAGRAM OF THE 3/4-WATT UNIT. It is almost identical to the 5-watt version except that it has no 5-watt stage or crystal-switching circuitry.

above 100 MHz. IF amplifier Q12 has a 20 dB gain, and it provides an IF signal to drive a

standard TV receiver tuned to VHF Channel 3 or 4.

If audio transmission is also

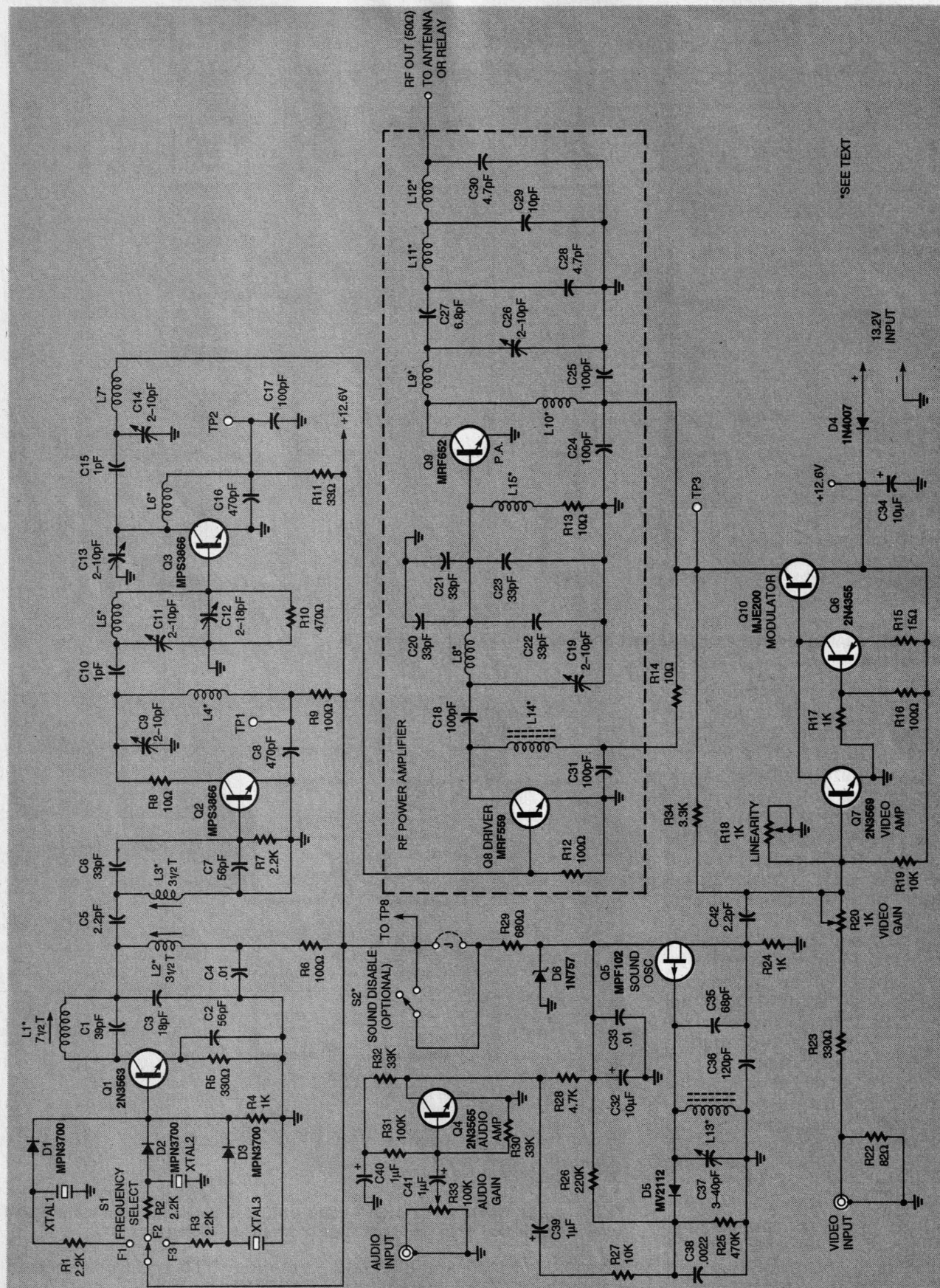


FIG. 3—5-WATT TRANSMITTER. A 30- to 40-mile range can be obtained with a suitable antenna and favorable height, terrain, local noise and interference levels, and path characteristics.

desired, Q5 and D6 generate a frequency-modulated audio subcarrier at 4.5 MHz. The 4.5 MHz subcarrier is fed to video amplifier Q6-Q7, and it modulates the transmitted carrier. An optional subcarrier switch mutes the sound if desired. This feature is also useful for test purposes. The FM subcarrier produces a "fuzz" on the monitor waveform from Q14 if not turned off. The subcarrier can also generate colored stripes on the video from Q14 because of a beat with the 3.58-MHz color subcarrier. This occurs because some monitors do not have a sound trap in their video input circuits.

The $\frac{3}{4}$ -watt version has a similar RF chain (see Fig. 2). Oscillator Q1 and doublers Q2 and Q3 are almost identical to those in the 5-watt version. Transistor Q4 serves as a power amplifier while Q5, Q6, and Q7 make up the video amplifiers and modulator, respectively. The main difference is in its lack of the 5-watt stage and crystal-switching circuitry.

A detailed circuit description of the 5 watt transceiver follows. Remember that the 5-watt transmitter is the transceiver without the downconverter, line sampler, and the RF relay. Also, much of the circuitry in the $\frac{3}{4}$ -watt version is identical to the circuitry in the 5-watt unit, so only the 5-watt unit will be discussed here.

ATV transmitter

Figure 3 is the schematic for the 5-watt transmitter, Fig. 4 is the schematic for the Mini ATV, and Fig. 5 is the schematic for the ATV Jr. Transistor Q1, a 2N3563, forms part of the crystal oscillator. Depending on which of three crystals is selected by S1, a +12-volt bias is applied to R1, R2, or R3, which forward biases PIN diode D1, D2, or D3. The bias causes the PIN diode to appear as a resistance of a few ohms or less. Resistor R4 completes the path to ground and forms the bottom leg of a voltage divider, supplying about +3 volts bias to Q1's base. The corresponding crystal is connected to the base of Q1

PARTS LIST—5-WATT TRANSMITTER

All resistors are $\frac{1}{8}$ -watt, 5%, unless otherwise specified.

R1-R3, R7—2200 ohms
R14—10 ohms, $\frac{1}{4}$ -watt
R4, R17—1000 ohms
R5, R23—330 ohms
R6, R9, R12, R16—100 ohms
R8, R13—10 ohms
R10—470 ohms
R11—33 ohms
R15—15 ohms
R18, R20—1000 ohms, potentiometer
R19, R27—10,000 ohms
R21—not used
R22—82 ohms
R25—470,000 ohms
R26—220,000 ohms
R28, R34—4700 ohms
R29—680 ohms
R30, R32—33,000 ohms
R31—100,000 ohms
R33—100,000 potentiometer
R34 (alternate value)—2200 or 3300 ohms

Capacitors

C1—39 pF, NPO
C2, C7—56 pF, NPO
C3—18 pF, NPO
C4, C33—0.01 μ F, disc GMV
C5, C42—2.2 pF, NPO (C42 alternates 1 or 3.3 pF)
C6—33 pF, NPO
C8, C16—470 pF, 20% disc
C9, C11, C13, C14, C19, C26—2–10 pF, trimmer (yellow body)
C10, C15—1 pF, NPO
C12—2–18 pF, trimmer (green body)
C17, C18, C24, C25, C31—100 pF, chip 60 $\times$ 120
C20—C23—33 pF, chip 60 $\times$ 120
C27—6.8 pF, NPO
C28, C30—4.7 pF, NPO
C29—10 pF, NPO
C32—10 μ F, 16 volts, electrolytic
C34—10 μ F, 16 volts, tantalum chip
C35—68 pF, NPO or SM
C36—120 pF, NPO or SM
C37—3–40 pF, trimmer (gray body)
C38—0.0022 μ F, 50 volts, Mylar
C39—C41—1 μ F, 35 or 50 volts, aluminum electrolytic

Semiconductors

D1-D3—Motorola MPN 3404 PIN diode (alternate MPN 3700)

D4—1N4007 diode
D5—Motorola MV2112 varactor diode
D6—1N757 diode
Q1—2N3563 transistor
Q2, Q3—Motorola MPS3866 transistor
Q4—2N3565 transistor
Q5—Motorola MPF102 FET
Q6—2N4355 transistor
Q7—2N3569 transistor
Q8—Motorola MRF559 transistor (alternate MRF627)
Q9—Motorola MRF652 transistor
Q10—Motorola MJE200 transistor
Inductors (all coils wound on 8-32 mandrel unless noted—inductances below 50 nH are approximate and may vary $\pm$ 10nH)
L1—125 to 300 nH ($7\frac{1}{2}$ turns No. 22 enameled with Cambion Blue 8-32 $\times$ $\frac{1}{4}$ slug)
L2, L3—50 to 100 nH ($3\frac{1}{2}$ turns No. 22 enameled with Cambion Blue 8-32 $\times$ $\frac{1}{4}$ slug)
L4—30 nH (4 turns No. 22 tinned)
L5—39 nH (5 turns No. 22 tinned)
L6, L8—5 nH ($\frac{1}{2}$ turn No. 22 tinned)
L7—10 nH ($\frac{1}{2}$ turns No. 22 tinned)
L9—7 nH ($\frac{1}{2}$ turn No. 20 tinned, 0.375" dia.)
L10—40 nH (5 turns No. 22 tinned)
L11—20 nH (2 turns No. 20 tinned)
L12—12 nH (1 turn No. 29 tinned)
L13—11 μ H (12 turns No. 22 enameled on 0.375" toroid core)
L14—Bead choke, 43 matl
L15—part of R13

Other components

XTAL1—54.90625 MHz crystal
XTAL2—54.25000 Mhz crystal
XTAL3—53.28125 MHz crystal
S1—SP3T switch and hardware
S2—SPST toggle switch

Miscellaneous: 2 RCA jacks, 1 female BNC, 1 power connector, 1 LED, 1 1000-ohm $\frac{1}{4}$ -watt resistor, No. 22 enameled wire, No. 20 tinned wire, No. 22 tinned wire, No. 32 enameled wire, heatsink material, 1 TO220 mica insulator, 5 No. 2 $\times$ $\frac{1}{4}$ " BHMS, 5 No. 2 nuts, 6 No. 4 locks, 1 No. 4 $\times$ $\frac{1}{2}$ " BHMS, 1 No. 4 nut, 1 No. 8 nut, 1 No. 8 lock, 1 No. 8 $\times$ 1" BHMS (for use as coil form), transmitter PC board, power-amp PC board

through a PIN diode. In the $\frac{3}{4}$ -watt version, R1 and R2 bias Q1's base to about +3 volts, and XTAL1 is connected between its base and ground.

At its series-resonant frequency, XTAL1 appears as a low impedance, effectively grounding the base of Q1. This forms a common-base oscillator at the crystal's series-resonant frequency. Frequency-selector switch S1 applies 12 volts to the selected crystal. Components L1

and C1 provide a high impedance load for the collector of Q1; L1 and C1 should be resonant at a frequency slightly higher than the crystal frequency (around 55 MHz or so). Because 55-MHz crystals are used, the oscillator frequency must be multiplied by 8 to obtain 440 MHz. While three doubler stages or a doubler and a quadrupler stage could do that, quadruplers are not usually efficient.

To minimize the number of

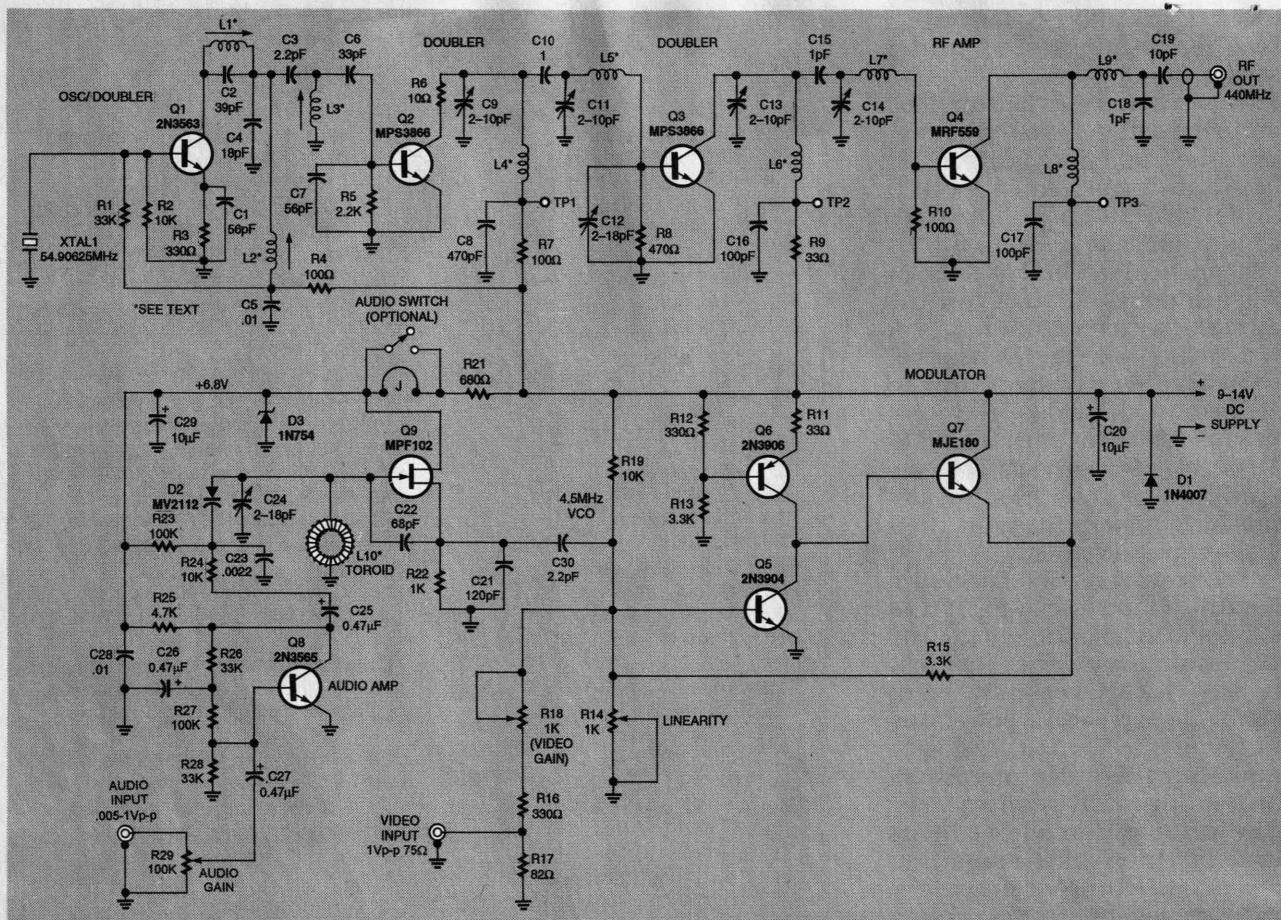


FIG. 4—MINI ATV. This $\frac{3}{4}$ -watt unit, complete with audio capability is useful where size is a factor.

stages, the oscillator is made to act as its own doubler. This is done by placing a resonant circuit—L2 and C3—in series with the collector circuit, tuned to twice the crystal frequency. Capacitor C4 acts as a bypass capacitor. In the $\frac{3}{4}$ -watt Mini ATV versions, the corresponding parts C4, L2, and C5 perform the same function. This circuit has a high impedance at 110 MHz, and it also appears as a very low impedance at 55 MHz. Similarly, L1-C1 appears as a high impedance at 55 MHz but a very low impedance at 110 MHz. Capacitor C5 couples 110 MHz energy to resonant circuit L3-C6-C7, which is tuned to 110 MHz. With this double-tuned circuit, a fairly clean 110-MHz signal is produced.

The RF voltage at the junction of C6, C7, and bias resistor R7 drives the base of Q2 (an MPS3866) fairly hard at 110 MHz. This results in a signal at the collector of Q2 that is rich in

harmonics. The collector of Q2 is connected to oscillation-suppressor resistor R8 and to resonant-circuit C9-L4-C8 tuned to 220 MHz. Components C13 and L6 select the 220-MHz component of the signal and couple it via C10 to another tuned circuit, C11-L5-C12, also tuned to 220 MHz.

Notice that test point TP1 is located at the "cold" side of L4 in both the 5- and $\frac{3}{4}$ -watt ver-

sions. It is the test point for measuring the DC current in the collector circuit of Q2. As more RF drive is applied to the base of Q2, more current is drawn by Q2. By monitoring the DC voltage drop across 100-ohms resistor R9 (R7 in the $\frac{3}{4}$ -watt versions), it is possible to tune L1, L2, and L3 properly without detuning them. Adjust L1 to produce a voltage drop across R9, and then adjust L2 and L3 to maximize the drop, which can be as much as 4 to 5 volts DC. Base resistor R10 supplies 220-MHz bias to Q3 after RF rectification.

Transistor Q3 is driven hard, causing it to act as a highly nonlinear amplifier. This stage draws 30 to 50 milliamperes of current. Tuned circuit C11-L5-C12 at the base of Q3 couples 220-MHz energy to Q3. Capacitor C10 couples the collector circuit of Q2 to the base of Q3. This provides a double-tuned circuit at 220 MHz, yielding good rejec-



5-WATT ATV TRANSCEIVER. This small package contains all the circuitry you'll need for transmitting, receiving, and downconverting a video signal complete with sound.

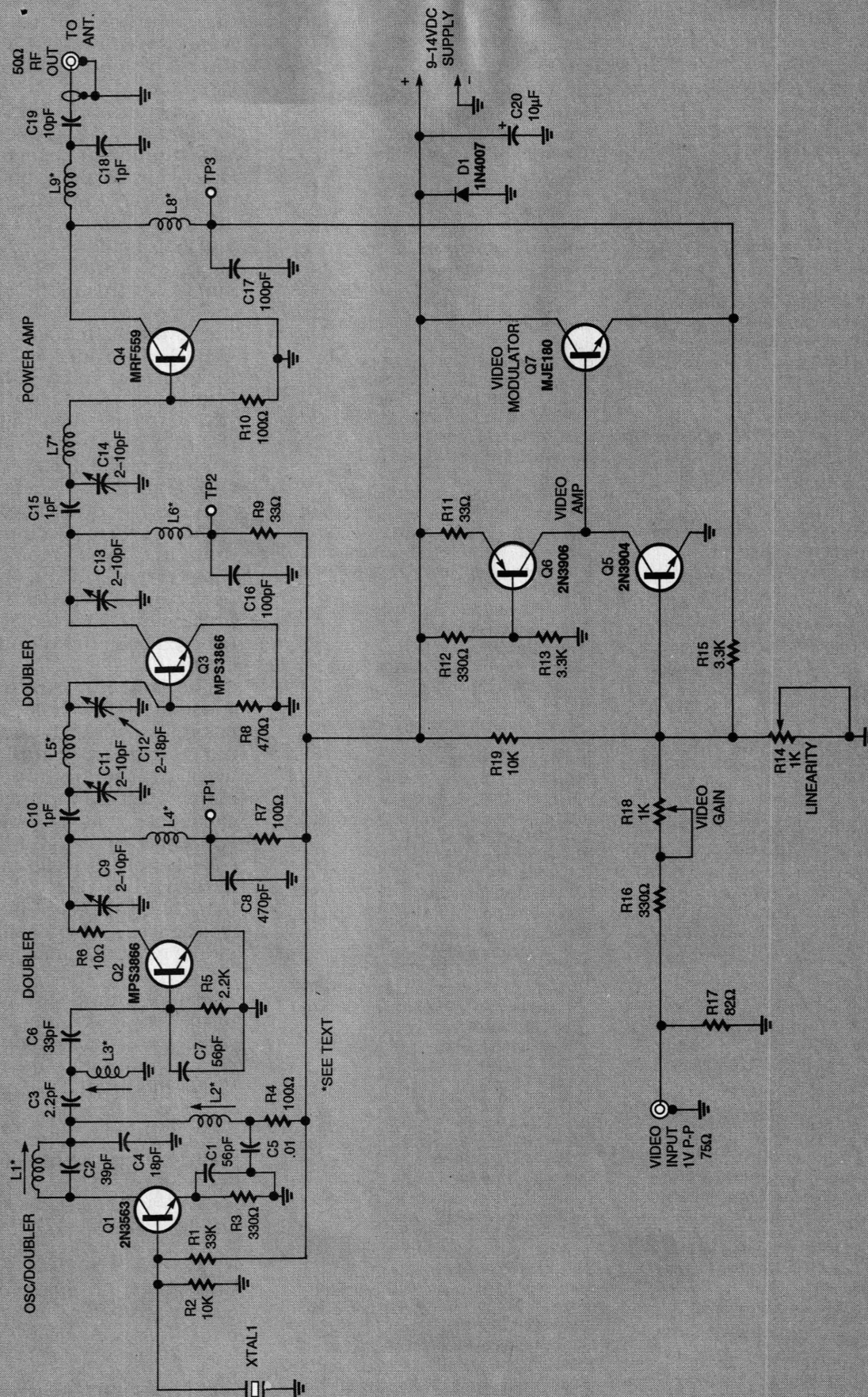


FIG. 5—ATV JR. If keeping cost as low as possible is of paramount concern, and audio output is not needed, build the ATV Jr.

PARTS LIST—5-WATT TRANSCEIVER

All resistors are 1/8-watt, 5%, unless specified

R1-R3, R7, R49—2200 ohms
R14—10 ohms, 1/4-watt
R4, R17, R53, R55—1000 ohms
R5, R23, R45, R47, R56—330 ohms
R6, R9, R12, R16, R35, R36—100 ohms
R8, R13—10 ohms
R10, R41—470 ohms
R11—33 ohms
R15, R44—15 ohms
R18, R20—1000 ohms, horizontal-mount potentiometer
R19, R27, R46—10,000 ohms
R21—not used
R22—82 ohms
R25, R39—470,000 ohms
R26—220,000 ohms
R28, R34, R42—4700 ohms (7× gain
R34 = 3300 ohms, 5X gain R34 = 2200 ohms)
R30, R32—33,000 ohms
R31, R38, R51—100,000 ohms
R33—100,000 ohms, horizontal-mount potentiometer
R29, R43—680 ohms
R37—47 ohms
R40—220 ohms
R48—6800 ohms
R50—1000 ohms, thumbwheel potentiometer
R52—470 ohms, 1/4-watt
R54—3300 ohms (4700 ohms for normal use)

Capacitors

C1, C53—39 pF, NPO
C2, C7—56 pF, NPO
C3, C52—18 pF, NPO
C4, C33, C53—C55, C62, C67, C69, C70—0.01 μF disc GMV
C5, C42—2.2 pF, NPO (alternate C42—1 or 3.3 pF)
C6, C56—33 pF, NPO
C8, C16—470 pF, 20% disc
C9, C11, C13, C14, C19, C26, C43, C46, C57, C59—2-10 pF trimmer (yellow body)
C10, C15, C60, C64—1 pF, NPO
C12—2-18 pF trimmer (green body)
C17, C18, C24, C25, C44, C45, C47—C49—100 pF, chip
C20—C23—33 pF, 60×120 chip
C27—6.8 pF, NPO
C28, C30—4.7 pF, NPO
C29, C65—10 pF, NPO
C32, C63, C66—10 μF, 16 volts, electrolytic
C34—10 μF, 16 volts, tantalum chip
C35—68 pF, NPO or SM
C36—120 pF, NPO or SM
C37—3-40 pF, trimmer (gray body)
C38—0.0022 μF, 50 volts, mylar
C39, C40, C41—1 μF, 35 or 50 volts, electrolytic
C50—0.6 pF (part of PC board)
C57—6.8 pF, NPO
C58—5.6 pF, NPO
C61—100 pF, NPO
C68—470 μF, 16 volts, electrolytic

Inductors (all coils wound on 8-32 mandrel unless noted—inductances

below 50 nH are approximate and may vary ±10nH)

L1—125 to 300 nH (7½ turns No. 22 enameled with Cambion Blue 8-32×¼ slug)
L2, L3—50 to 100 nH (3½ turns No. 22 enameled with Cambion Blue 8-32×¼ slug)
L4—30 nH (4 turns No. 22 tinned)
L5—39 nH (5 turns No. 22 tinned)
L6, L8—5 nH (½ turn No. 22 tinned)
L7—10 nH (1½ turns No. 22 tinned)
L9—7 nH (½ turn No. 20 tinned, 0.375" dia.)
L10—40 nH (5 turns No. 22 tinned)
L11—20 nH (2 turns No. 20 tinned)
L12—12 nH (1 turn No. 29 tinned)
L13—11 μH (12 turns No. 22 enameled on 0.375" toroid core)
L14—Bead choke, 43 matl
L15—part of R13
L16—7 nH (½ turn No. 20 tinned)
L17, L18—20 nH (2 turns No. 20 tinned)
L19—75 nH (5 turns No. 22 enameled)
L20—200 to 550 nH (11½ turns No. 22 enameled with Cambion Blue, 8-32×¼ slug)
L21—8 nH (½ turn No. 20 square loop)
L22—18 μH RF choke

Semiconductors

D1-D3—Motorola MPN 3404 PIN diode (alternate MPN 3700)
D4—1N4007 diode
D5—Motorola MV2112 varactor diode
D6, D10—1N757 diode
D7, D8, D11—8200-2835 diode
D9—Motorola MV2103 varactor diode
Q1, Q12—2N3563 transistor
Q2, Q3—Motorola MPS3866 transistor
Q4—2N3565 transistor
Q5—Motorola MPF102 FET
Q6—2N4355 transistor
Q7—2N3569 transistor
Q8—Motorola MRF559 transistor (alternate MRF627)
Q9—Motorola MRF652 transistor
Q10—Motorola MJE200 transistor
Q11—NEC 25137 or NEC 25139 FET
Q13—Motorola MPSH81 transistor
Q14—2N3904 transistor

Other components

XTAL1—54.90625 MHz
XTAL2—54.25000 MHz
XTAL3—53.28125 MHz
S1—2P3T switch
S2—SPST toggle switch
RY1—12-volt DIP relay
M1—SBL-1 mixer

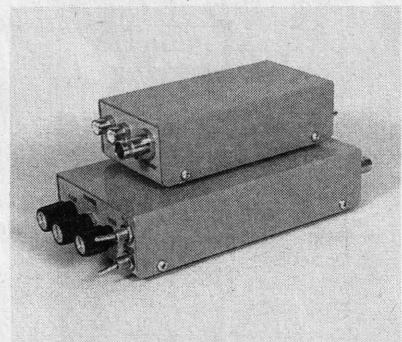
Miscellaneous: 2 RCA jacks, 1 female BNC, 1 power connector, 1 LED, 1 1000-ohm 1/4-watt resistor, No. 22 enameled wire, No. 20 tinned wire, No. 22 tinned wire, No. 32 enameled wire, heatsink material, 1 TO220 mica insulator, 5 No. 2×¼" BHMS, 5 No. 2 nuts, 1 12" teflon cable, 1 2" potentiometer shaft, 6 No. 4 locks, 1 No. 4×½" BHMS, 1 No. 4 nut, 1 No. 8 nut, 1 No. 8 lock, 1 8×1" BHMS (for use as coil form), transceiver PC Board, power-amp PC board

tion of unwanted frequencies. Adjust trimmer C12 for optimum matching to the base of Q3.

Capacitors C16 and C17 form a resonant circuit tuned to 440 MHz. Inductor L6 is a half-turn loop of No. 20 wire (about 0.006 microhenry). Capacitor C17 has a low impedance at 440 MHz, and C15 couples the 440-MHz energy to tuned circuits C14 and L7. At this point, about 50 to 75 milliwatts of RF energy at the transmit frequency (426 to 440 MHz) is available. The same RF exciter circuitry (Q1-Q2-Q3) is in both the 5- and 3/4-watt transmitters. Test point TP2 permits monitoring collector current of Q3 for tuning C9, C13, and C12. It is adjusted by peaking the three trimmer capacitors for a maximum current drain in Q3's collector. In the 5-watt version, Q8 and Q9 form a power amplifier. Resistor R12 provides self-bias for Q8, which receives base drive from tuned circuit C12-L7.

The collector circuit of Q8 consists of RF choke L14, bypass capacitor C31, and matching network C19 and L8. All bypass capacitors are ceramic chips. They are the only bypass capacitors that are both small and effective at 440 MHz. The collector of Q9 feeds matching network L9, C26, C27, C28, L11, C29, L12, and C30, which also acts as a lowpass filter. Harmonic output is a low 45 to 50 dB.

Transistors Q8 and Q9, which form the power amplifier assembly, are mounted on a separate heatsinked subassembly. Video-modulated DC is fed to



5-WATT TRANSMITTER (bottom) has audio output but it can't receive or down-convert video. **Mini ATV (top)** is a 3/4-watt unit that can transmit audio and video.

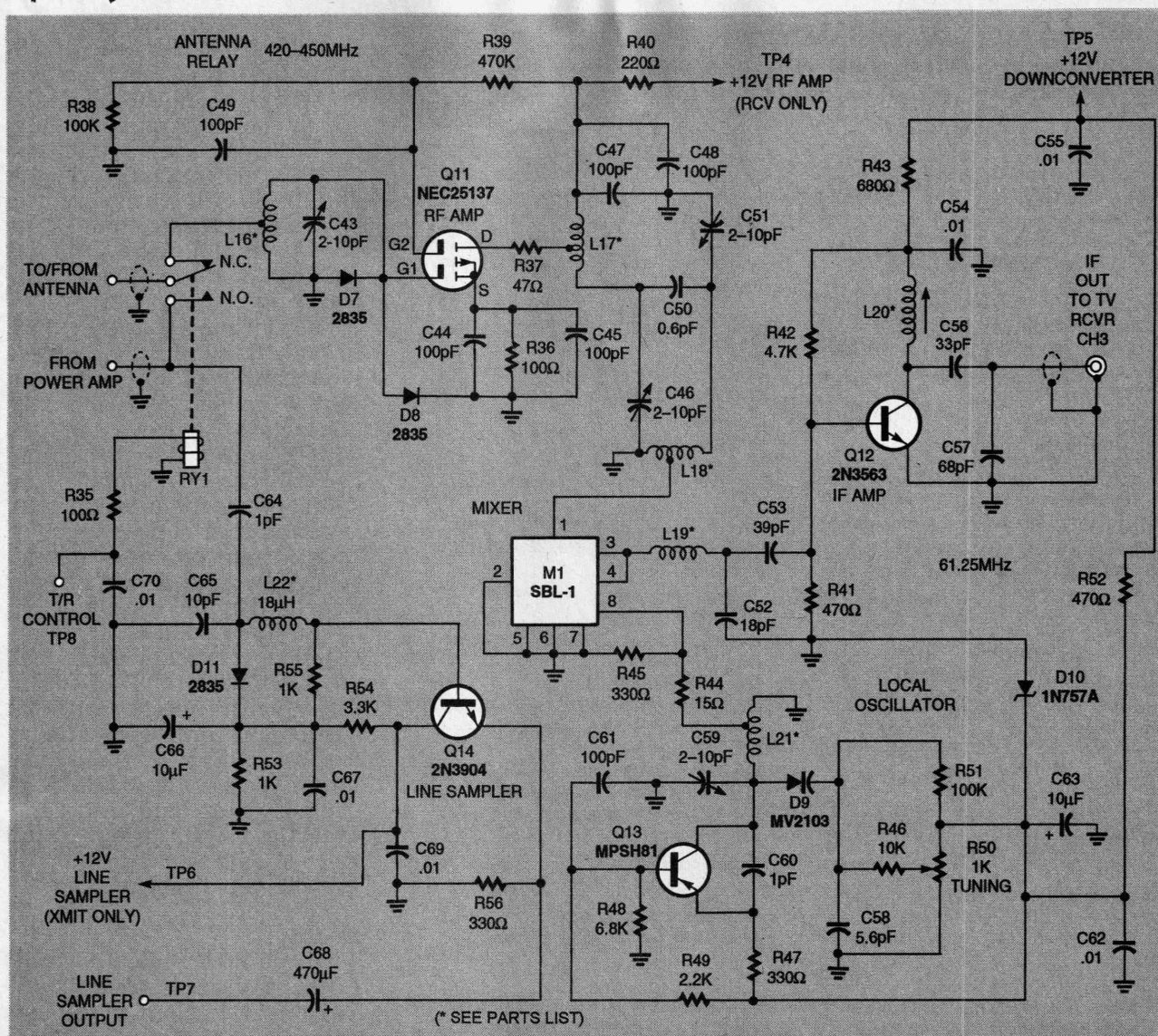


FIG. 6—DOWNCONVERTER, antenna relay, and line sampler. The downconverter consists of a tuned RF stage with a low-noise, dual-gate GASFET, a double-balanced diode mixer, an IF amplifier, and a varactor-tuned local oscillator.

PARTS LIST—ATV JR.

All resistors are 1/8-watt unless otherwise specified

R1—33,000 ohms
R2, R19—10,000 ohms
R3, R12, R16—330 ohms
R4, R7, R10—100 ohms
R5—2200 ohms
R6—10 ohms
R8—470 ohms
R9, R11—33 ohms
R13—3300 ohms
R14, R18—1000 ohms, horizontal-mount potentiometer
R15—3300 ohms (alternate 2200 to 4700 ohms)
R17—82 ohms

Capacitors

C1, C7—56 pF, NPO
C2—39 pF, NPO
C3—2.2 pF, NPO

C4—18 pF, NPO
C5—0.01 μF, disc GMV
C6—33 pF, NPO
C8—470 pF, disc GMV
C9, C11, C13, C14 2–10 pF trimmer (yellow body)
C10, C15, C18—1 pF, NPO
C12—2–18 pF trimmer (green body)
C16, C17—100 pF, chip
C19—10 pF, NPO
C20—10 μF, 16 volts, chip

Semiconductors

D1—1N4007 diode
D2—Motorola MV2112 varactor diode
Q1—2N3563 transistor
Q2, Q3—Motorola MPS3866 transistor
Q4—Motorola MRF559 transistor
Q5—2N3904 transistor
Q6—2N3906 transistor
Q7—Motorola MJE180 transistor

Inductors (all coils wound on 8-32 × 1/4" form unless noted—inductances below 50 nH are approximate and may vary ± 10 nH)

L1—125 to 300 nH (7 1/2 turns No. 22 enameled with 8-32 × 1/4" Cambion Blue slug)
L2, L3—50 to 100 nH (3 1/2 turns No. 22 enameled with 8-32 × 1/4" Cambion Blue slug)
L4—30 nH (4 turns No. 22 tinned)
L5, L8—39 nH (5 turns No. 22 tinned)
L6—5 nH (1/2 turn No. 22 tinned 0.375" dia.)
L9—25 nH (2 1/2 turns No. 22 tinned)

Other components

XTAL1—54.90625 MHz crystal
Miscellaneous: ATV Jr. PC board, 8-32 screw for coil winding

PARTS LIST—MINI ATV

All resistors are 1/8-watt unless otherwise specified

R1, R26, R28—33,000 ohms
R2, R19, R24—10,000 ohms
R3, R12, R16—330 ohms
R4, R7, R10—100 ohms
R5—2200 ohms
R6—10 ohms
R8—470 ohms
R9, R11—33 ohms
R13—3300 ohms
R14, R18—1000 ohms, thumbwheel trimmer potentiometer
R15—3300 ohms (alternate 2200 to 4700 ohms)
R17—82 ohms
R20—not used
R21—680 ohms
R22—1000 ohms
R23, R27—100,000 ohms
R25—4700 ohms
R29—100,000 thumbwheel trimmer potentiometer

Capacitors

C1, C7—56 pF, NPO
C2—39 pF, NPO
C3—2.2 pF, NPO
C4—18 pF, NPO
C5, C28—0.01 μ F, Disc GMV
C6—33 pF, NPO
C8—470 pF, Disc GMV
C9, C11, C13, C14—2–10 pF trimmer (yellow body)
C10, C15, C18—1 pF, NPO
C12, C24—2–18 pF trimmer (green body)
C16, C17—100 pF, chip
C19—10 pF, NPO
C20—10 μ F, 16 volts, chip
C21—120 pF, NPO or SM
C22—68 pF, NPO

C23—0.0022 μ F, 50 volts, mylar
C25—C27—0.47 or 1.0 μ F, 35 volts, tantalum electrolytic
C29—10 μ F, 16 volts, electrolytic
C30—1 to 3.3 pF, NPO (1 pF, and 2.2 pF supplied in Mini-ATV kits—value determines sound subcarrier level)

Inductors (all coils wound on 8-32 $\times$ 1/4" form unless noted—inductances below 50 nH are approximate and may vary $\pm$ 10 nH)

L1—125 to 300 nH (7 1/2 turns No. 22 enameled with 8-32 $\times$ 1/4" Cambion Blue slug)
L2, L3—50 to 100 nH (3 1/2 turns No. 22 enameled with 8-32 $\times$ 1/4" Cambion Blue slug)
L4—30 nH (4 turns No. 22 tinned)
L5, L8—39 nH (5 turns No. 22 tinned)
L6—5 nH (1/2 turn No. 22 tinned 0.375" dia.)
L9—25 nH (2 1/2 turns No. 22 tinned)
L10—11 μ H (12 turns No. 22 enameled on toroid)

Semiconductors

D1—1N4007 diode
D2—Motorola MV2112 varactor diode
D3—1N754 diode
Q1—2N3563 transistor
Q2, Q3—Motorola MPS3866 transistor
Q4—Motorola MRF559 transistor
Q5—2N3904 transistor
Q6—2N3906 transistor
Q7—Motorola MJE180 transistor
Q8—2N3565 transistor
Q9—Motorola MPF102 FET

Other components

XTAL1—54.90625 MHz crystal

Miscellaneous: Mini ATV PC board, 8-32 screw for coil winding form

Note: The following items are available from North Country Radio, PO Box 53, Wykagyl Station, New Rochelle, NY 10804-0053:

- 5-watt transceiver kit (contains PC boards, all parts that mount on them, chassis connectors suitable for basic operation, and three crystals for 439.25, 434.0, and 426.25 MHz)—\$179.00
- 5-watt transmitter kit (contains PC boards, all parts that mount on them, suitable chassis connectors for basic operation, and three crystals for 439.25, 434.0, and 426.25 MHz—downconverter and line sampler components and RF switching relay NOT included)—\$149.00
- 3/4-Watt Mini-ATV kit (contains PC board and all parts that mount on it with crystal for 439.25 MHz operation)—\$79.00

- 3/4-Watt ATV Jr. kit (contains PC board and all parts that mount on it and crystal for 439.25 MHz operation)—\$59.00
- Test Crystals (CH14, 15, 16, 17, 18, for test or export only—not legal for on the air transmission in the USA) and others for 434.0-, 427.75-, 426.25-, and 421.25-MHz are available—\$8.50 each, specify channel
- Other ATV kits for 440- and 915-MHz and CCD cameras are available—contact North Country Radio for details
- A complete catalog of kits is available from North Country Radio—send \$1.00 with a self addressed stamped (52 cents) envelope Please include \$4.50 for the first item and add \$1.00 for each additional item for postage and handling. New York residents add sales tax.

L10 and through R14 to L14. Capacitors C24, C25, and C31 have low impedance at 440 MHz but high impedance at the higher video frequencies. The modulator must deliver video at 10 to

12 volts p-p into a 12-ohm load. This requires a power amplifier with a response from DC to 4 MHz. Modulator Q10 is installed on the same heatsink as Q8 and Q9.

A 440-MHz signal is fed through L7 to the base of Q8. About 0.5 to 1 watt of RF is produced by Q4, depending on the supply voltage. A matching network is formed by L9, C18, and C19. The network is broadband, and fixed tuning was found to be adequate. Notice that in the 3/4-watt version, the V_{CC} supply fed to Q4 is the only modulating source.

The video modulating circuit in both transmitter circuits is identical except for component values. Transistor Q10 is part of a feedback-pair video amplifier with Q7 acting as a video amplifier. Transistor Q6 supplies constant current to the base of Q10, assuring drive at peak voltages. This current produces a larger voltage swing than if a resistor were used in the collector of Q6.

The quiescent point (zero signal voltage) is set by the ratio of feedback-resistor R34 to linearity-control resistance R18, as well as gain-control potentiometer R20. Adjust R20 for maximum gain without white or black clipping, and adjust R18 to set the operating point. Input video should be between 0.7 to 1.5 volts p-p, negative sync.

Audio

The audio channel (not used in the ATV Jr.) is basically the same in all versions. It consists of a preamplifier stage built around a 2N3565 (Q4) with a voltage gain of about +43 dB. In both 5-watt boards, audio is fed to gain-control R33 through coupling capacitor C41 to the base of Q4. Audio developed across R28 is coupled by C39 and R27 to varactor diode D5. The oscillator frequency is determined by L13 and the capacitance that shunts it. This is the series equivalent of C36 and C35, the input capacitance of Q5, trimmer C37, and the varactor diode capacitance.

Adjust the oscillator by setting C37 for operation at 4.5 MHz. When an audio signal is applied to the junction of C38, R25, R26 and D5, the effective capacitance of D5 varies with the instantaneous voltage across it. This response causes

Continued on page 84

AMATEUR TV STATION

continued from page 40

the oscillator frequency to change about 20 kHz per volt of applied signal. About 2 to 2.5 volts p-p is required for 25-kHz deviation of the 4.5-MHz subcarrier. About 6 to 7 millivolts of audio signal is required at the base of Q4 to obtain full deviation (25 kHz).

The FM 4.5-MHz subcarrier is taken from the source of Q5 through C42. The value of C42 sets the subcarrier level; C42 also removes the DC level present at this point. If it were not removed, it might change the effective Q point of the video amplifier. The audio subcarrier is fed to the video amplifier Q7 where it is mixed with the video signal. This design makes a separate sound transmitter/RF system unnecessary.

If desired, optional switch S2 can be inserted in series with R29 to disable the audio subcarrier. This is often done in ATV work, where a separate audio link (such as a 2-meter FM) might be used instead. With this modification, it will not be necessary to turn on the video transmitter to talk, and it is useful during test and setup procedures. If S2 is not needed, install the jumper across the switch pads on the PC board.

The Mini ATV has the same audio circuitry as the 5-watt transmitter circuit with one exception. The Mini-ATV requires 9 volts to operate, so the 9-volt regulator D6 in the 5-watt version is replaced with a 6-volt diode (D3) in the Mini-ATV. Also, the varactor diode (D2 in the Mini ATV) is biased to the full 6-volt regulated voltage, through R23. The ATV Jr., of course, has no audio channel so there are no audio components on the PC board.

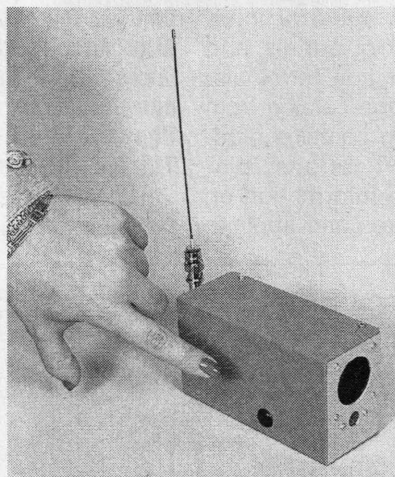
Downconverter

The downconverter, antenna relay, and line sampler section of the 5-watt transceiver are shown in Fig. 6. If you are building only a transmitter, you can skip this section. The downconverter consists of a tuned RF

stage with a low-noise, dual-gate GASFET, a double-balanced diode mixer, an IF amplifier, and a varactor-tuned local oscillator.

Signals from the antenna pass through relay RY1 to a tap on antenna coil L16. The coil is tuned to signal frequency (420 to 440 MHz) by trimmer C43. Diodes D7 and D8 protect the gate of RF amplifier Q11 from excessive RF levels that might damage it. Transistor Q11 is biased to about 10 milliamperes drain current.

Capacitor C50, a value of about 0.6 picofarads, built into the PC board layout, couples RF into the tuned circuit consisting of C51 and L18. A tap on L18



THIS ATV UNIT contains a built-in CCD camera module and rechargeable battery pack. It's a hand-held wireless TV transmitter in itself.

can match the input impedance of double-balanced mixer M1 to the tuned circuit. Mixer M1 is fed with a local-oscillator signal of about +7 to +10 dBm (500 to 700 millivolts). Oscillator transistor Q13 is biased by R47, R48, and R49. The collector circuit consists of oscillator inductor L21, trimmer capacitor C57, and varactor diode D9, in series with 5.6-picofarad capacitor C58. The DC bias is applied via R51, R46, and potentiometer R50.

The setting of R50 determines the bias on D9 (from +1 to +9V). It tunes the LO frequency from 350 to 380 MHz; this will provide 60 to 72-MHz IF frequencies over the input signal range of 420 to 445 MHz.

This corresponds to channel 3 or channel 4 (VHF TV) so that a standard TV receiver can serve as an IF amplifier and display monitor for the transceiver.

Components D10, R52, C62 and C63 provide a regulated 9-volt DC source for the oscillator, and C60 provides feedback. The LO signal is fed to mixer M1 through R44 and R45 from a tap on L21. The IF signal is taken from pins 3 and 4 of M1 through lowpass filter L19 and C52, and coupling capacitor C53. The IF amplifier Q12 is biased by R43, R42, and R41. Coil L20, C56, and C57 are resonant at the IF frequency, and they match the collector circuit of Q12 to 50 to 75-ohm loads.

The downconverter has separate supply lines for the RF amplifier (TP4) and the mixer-LO-IF section (TP5). During transmit, V_{CC} is removed from the RF amplifier (Q11 and associated components), but the mixer-LO-IF is still functional. This allows the operator to pick up some transmitted signal with the downconverter and see it on the receiver—a handy feature for verifying transmission. The downconverter gain is cut from about 33 dB in its receive mode to about a 10 to 20 dB loss. If the downconverter feature is not needed and you want to disable it completely, the downconverter mixer-LO-IF lead (TP5) can be fed from +12-volts in the receive-only mode, as in the RF amplifier.

To monitor transmission quality, a line sampler circuit is included. This circuit samples the transmitted RF signal, demodulates it, and feeds the detected (demodulated) video to an emitter-follower output buffer and then to an output jack. Refer to Fig. 6. Capacitors C64 and C65 form a capacitive voltage divider across the RF power amplifier output line. About 1.5 volts of RF signal is available from the divider. Diode D11 is an envelope detector that produces a negative video signal. The video is fed through RF choke L22 to the emitter of Q14. The detected video appears across R56.

The base bias of Q14 is set at

about +3 volts through resistors R54 and R53 and bypass capacitors C66 and C67. Because Q14 is an emitter-follower, a low impedance video output is produced. Capacitor C68 couples video to the line sampler output jack. Sufficient output (about 1.5 volts) is available to drive most monitors. If more or less output is desired, change C65 to 6.8 or 15 picofarads, respectively. If a variable output is needed, substitute a trimmer with a value of 5 to 20 picofarads in place of C65. (This change was not found to be necessary in the prototypes.)

Relay RY1, a DIP reed relay, is activated through R35 and bypass C42 by voltage from the transmit supply lead. The relay has about 0.7 to 0.8 dB loss, but switching RF at 440 MHz at a reasonable cost is not easy. The relay can be omitted, and the RF output from the transmitter and the RF input to the down-converter could be brought out to separate jacks.

Next month

That's all for this month. If you want to build any ATV unit, collect all the parts you'll need for the unit you want to build. Next month's article will tell you how to build the ATV circuits. Ω

AUDIO UPDATE

continued from page 79

reproduced sound whose recorded acoustics are at odds with the acoustics of the room in which it is being reproduced. It just doesn't sound live. The sound is a lot more realistic when we are listening binaurally through headphones—which means no listening-room acoustics—when the listening-room acoustics are swamped out by multiple channels, or when the recording is played back in the same room it was made.

In any case, listen to some live, minimally-amplified performances against the best recorded material at hand. What differences do you hear? I would be interested to know if you think my SVA theory holds water. Ω

WHAT'S NEWS

continued from page 6

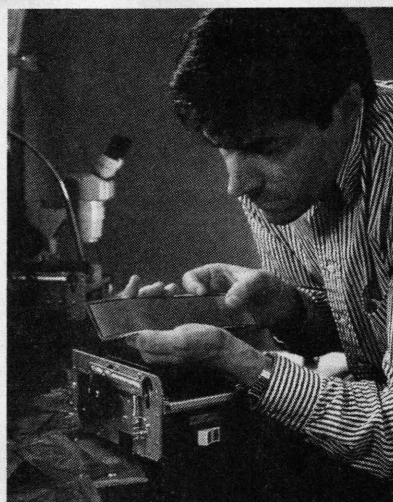


FIG. 2—PROTOTYPE, PLASTIC, rechargeable lithium-ion battery is held by Bellcore scientist Jean-Marie Tarascon.

weigh half as much. Moreover, they say lithium-ion technology poses none of the environmental problems of the most commonly used rechargeable cells and batteries based on nickel-cadmium or lead-acid chemistry.

"This is the first plastic, solid-state rechargeable battery," says Jean-Marie Tarascon, leader of Bellcore's battery team. "It does not contain such toxic metals as lead, cadmium, mercury, or cobalt.

What's more, no liquid will leak out if the battery is cut or punctured; this makes it safer to install and use." The new power cell is a refinement of one announced by Bellcore in late 1992, which had a liquid electrolyte.

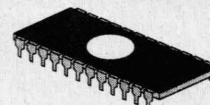
The new solid form of lithium-ion cell consists of five active layers encased in a metallized plastic bag that keeps out contaminants. Reading down from the positive side they are an aluminum mesh, a plastic anode layer containing manganese oxide (the positive pole), and then a layer of plastic that appears solid. It is actually a porous plastic structure impregnated with liquid electrolyte. Below that is the plastic cathode layer containing carbon (the negative pole), and at the bottom is copper mesh. The manufacturing process calls for fusing the layers together with heat.

When the power cell is being charged, lithium ions from the anode pass through the electrolyte layer and collect on the cathode. The reaction is reversed when current is being drawn from the cell: current flows across the terminals in the opposite direction.

According to Bellcore, their lithium-ion cell has twice the energy density of a nickel-cadmium (Ni-Cd) cell and 40% more than a nickel-metal hydride (Ni-Mhd) cell. Each cell produces a nominal voltage of

Continued on page 89

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CIRCLE 183 ON FREE INFORMATION CARD

BAY MARSTON

THIS ARTICLE CONTINUES THE Relay switching circuit theme of last month by presenting a selection of practical control circuits for electromagnetic relays. Last month's article discussed the basic operating principles and application rules for four well-known relay classes—the electromagnetic relay (EMR), the reed relay, the solid-state relay (SSR), and the CMOS bilateral switch.

High-impedance relays.

Electromagnetic and reed relays are low-impedance, electromechanical devices whose typical coil impedance values range from tens to hundreds of ohms while input drive signal sources typically have much higher impedance values. Maximum power is transferred between the input drive signal and the relay coil when the source and load impedances are matched.

This is accomplished by inserting additional circuitry between the input drive signal and the relay coil. These circuits can be either a transistor stage or a transistor stage with one or more IC logic gates. Figures 1 to 4 show four different ways to boost the input impedance of a relay with transistors and logic gates.

Transistor Q1 in Fig. 1 is organized as a simple common-emitter amplifier that increases the effective sensitivity of the 12-volt coil of relay RY1 about one hundred times (e.g., the current gain of Q1 in this circuit is about 100). The introduction of this amplifier stage reduces

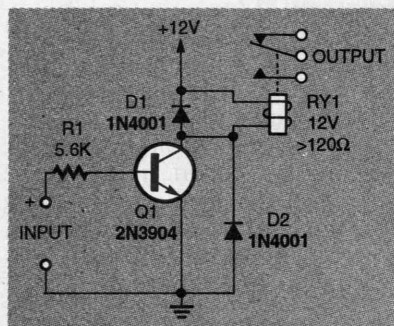
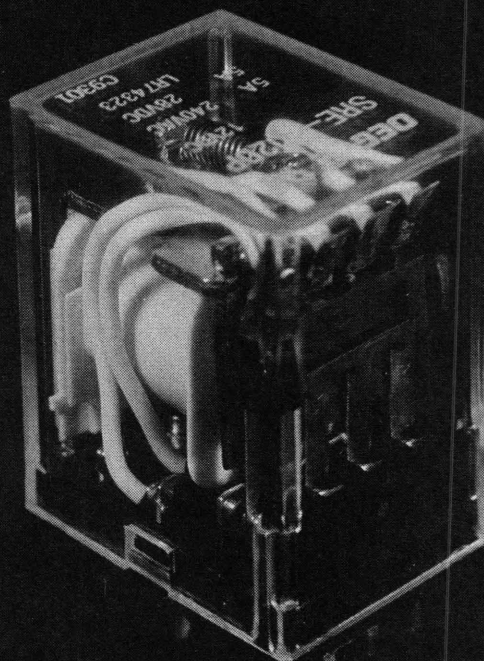


FIG. 1—NON-LATCHING TRANSISTOR-driven relay switching circuit.

RELAY OUTPUT



CIRCUITS

Learn to design transistorized circuits that can switch and time relay functions remotely with low-power input.

relay sensitivity to a few volts.

Resistor R1 puts safe limits on the input current to Q1, and it also determines the effective input impedance of the circuit. (The impedance is equal to the value of R1 plus about 1 kilohm.) Diodes D1 and D2 damp the coil's back EMF. (See last month's article.) The contacts of relay RY1 can control external circuitry.

Figure 1 is a schematic for a *non-latching* relay driver circuit. Relay RY1 is OFF when the input voltage is less than 600 millivolts and ON when the input voltage exceeds a few volts. This driver circuit can be made *self-latching* by modifying it as

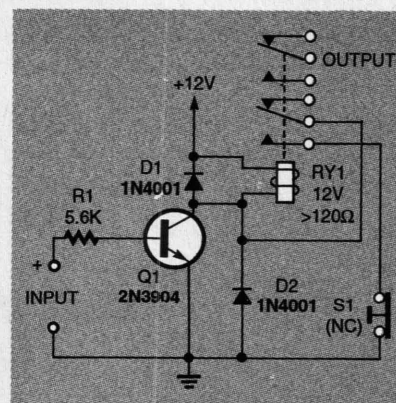


FIG. 2—SELF-LATCHING TRANSISTOR-driven relay switching circuit.

shown in Fig 2. Relay RY1 now has two sets of normally open (NO) contacts; the upper set is

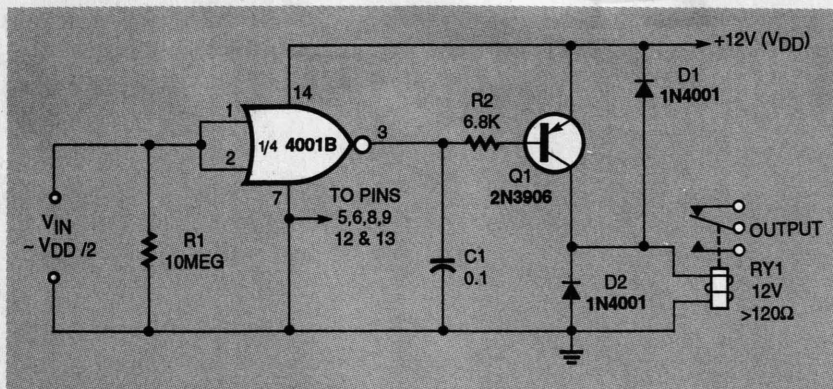


FIG. 3—HIGH-IMPEDANCE RELAY switching circuit.

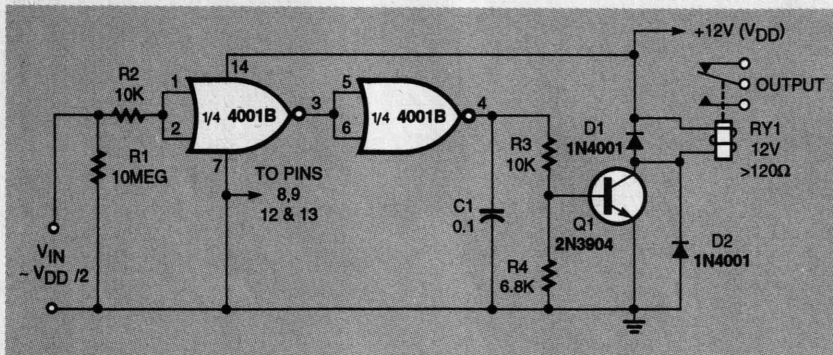


FIG. 4—MODIFIED HIGH-IMPEDANCE relay switching circuit.

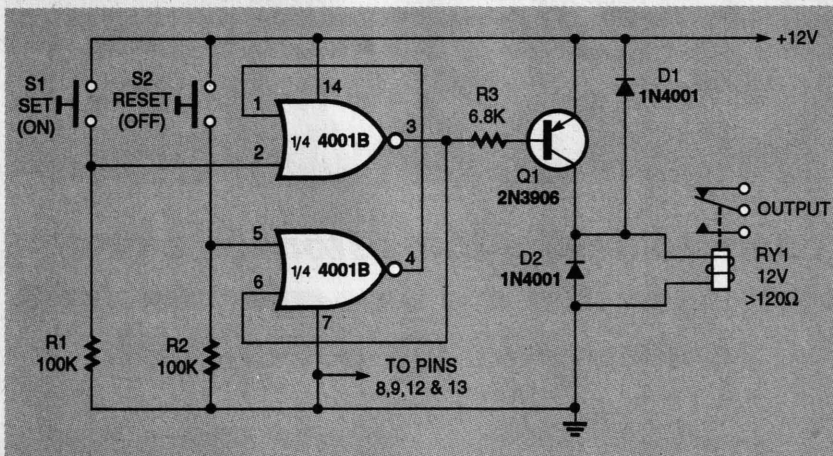


FIG. 5—BISTABLE RELAY SWITCHING circuit.

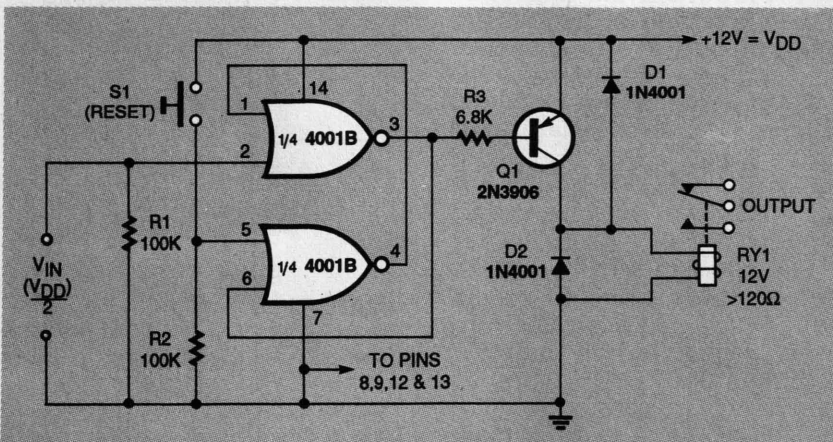


FIG. 6—SELF-LATCHING, HIGH-impedance relay switching circuit.

for the relay's output and the lower set is for latching. The latching set is in parallel with Q1, and it is actuated by normally closed (NC) switch S1.

As a result, these contacts bypass Q1 and self-latch RY1 after the relay has been initially energized. Once it has self-latched, the relay can be turned off again only by opening S1 or breaking the power connection.

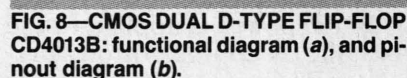
The circuits of Figs. 1 and 2 have input impedances of only a few thousand ohms. If desired, the input impedance can be raised to 10 megohms or more by driving Q1 through a CMOS buffer stage as shown in Figs. 3 and 4. In these circuits, the CMOS buffer consists of one of four NOR gates of a CD4001B quad, 2-input NOR gate. The gate is wired as an inverter by short-circuiting its two input pins 1 and 2 together.

Figure 3 has only a single CMOS inverter stage. Consequently, to ensure that the relay turns off when the input voltage is low, transistor driver Q1 must be a PNP device, in this instance a 2N3906. In Fig. 4, two CMOS inverter stages are wired in series, so that the overall signal inversion is zero. Consequently, Q1 can be an NPN transistor such as the 2N3904.

It is worth noting in both Figs. 3 and 4 that the input impedance actually equals the value of resistor R1, and that the relay actually turns ON (or OFF) when the input signal exceeds (or drops below) about one half the supply voltage. This is the transition voltage for the CD4001B CMOS input gate. At that value the gate operates in the linear mode. Capacitor C1 inhibits any high-frequency or transient signals that might appear at the gate when the circuit is operated in the linear mode.

Bistable circuits

A relay can be organized for bistable operation. In this mode it turns on and self-latches when a SET pushbutton switch is pressed, and it can be turned off again only by pressing a RESET pushbutton switch. Figure 5 is the schematic for a bistable relay switching circuit.



The relay turns on and self-latches when the input voltage rises above the transition value; the relay can then be turned off only by removing or reducing the input voltage and pressing

Capacitor C1 then discharges slowly through R2 when the switch is re-opened. This response eliminates false triggering caused by switch bounce or



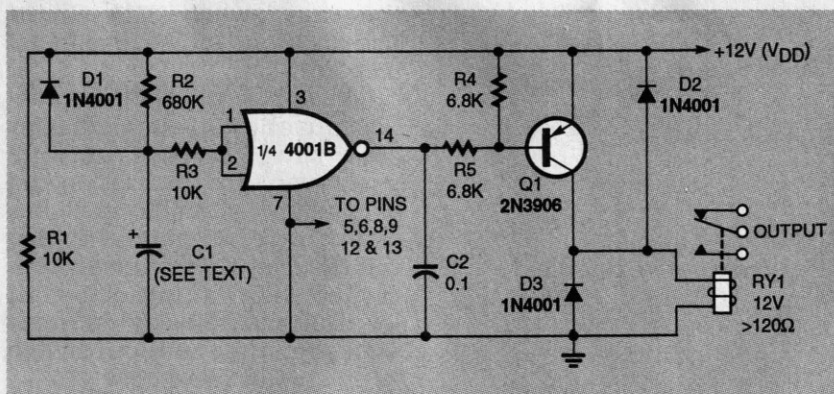


FIG. 10—DELAYED-TURN-ON RELAY switching circuit.

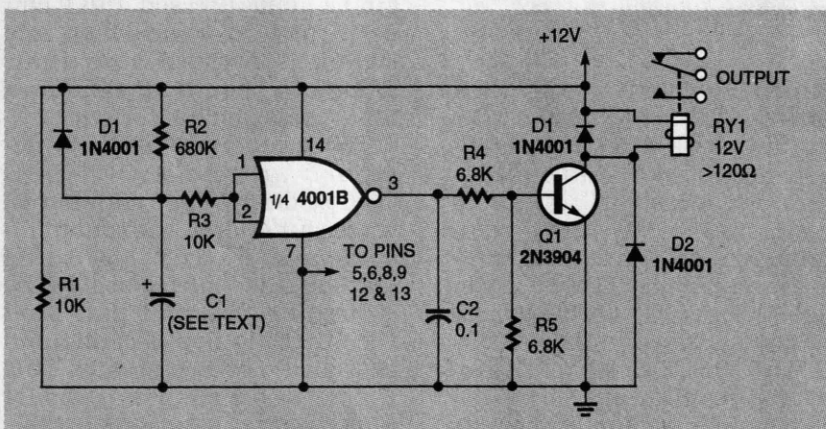


FIG. 11—AUTOMATIC TURN-OFF RELAY switching circuit.

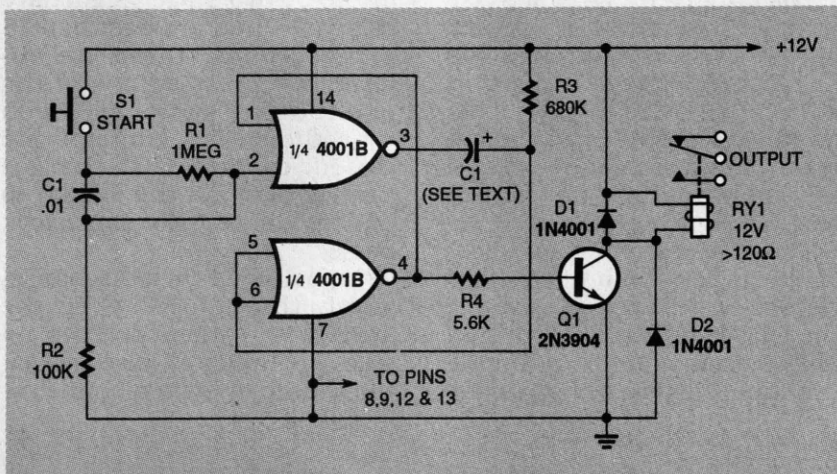


FIG. 12—ONE-SHOT MULTIVIBRATOR TIMER relay switching circuit.

chatter. As a result, transistor Q1 and the relay reliably change state each time a pushbutton switch is pressed.

The circuit illustrated in Fig. 7 will permit the relay to be actuated from many different locations. If this circuit is installed for use in a home or office it is recommended that approved switches suitable for use 120-volt AC operation be installed. Hall, landing, or cor-

ridor lights can be controlled from several different locations. Approved insulated twin-lead cable connecting the components can be concealed in the walls or along the baseboards.

Figure 8 shows the functional and pinout diagrams for the CMOS, dual D-type flip-flop. It consists of two identical flip-flops. Each flip-flop has independent DATA, SET, and CLOCK inputs and Q and $\bar{Q}$ outputs.

In the light control application, the circuit should be powered from the 120-volt AC line. Figure 9 shows how the circuit in Fig. 7 can be modified to operate from the AC line. Transformer T1 must be selected to provide a 24-volt, center-tapped output at 100 milliamperes or greater.

Timer circuits

Relays can be put to work in a wide variety of timer or time-delayed-switching applications with time delays from a fraction of a second to tens of hours. Figures 10 to 15 are practical examples of timing circuits for relay control.

Figures 10 to 12 show how CD4001B CMOS NOR gate ICs can produce time delays of up to several minutes with reasonable accuracy. The circuit in Fig. 10 offers delayed turn-on relay switching, and it operates as follows:

The CMOS gate is configured as a simple digital inverter. Its output is fed to the base of PNP transistor Q1 at the junction of resistor R5 and capacitor C2. The input to IC1 is taken from the junction of the time-controlled potential divider formed by resistor R2 and capacitor C1.

Before power is applied to the circuit, capacitor C1 is fully discharged. Therefore, the inverter input is grounded, and its output equals the positive supply voltage; transistor Q1 and relay RY1 are both off under this circuit condition.

When power is applied to the circuit, C1 charges through resistor R2, and the exponentially rising voltage is applied to the input of the CMOS inverter gate. After a time delay determined by the RC time constant values of capacitor C1 and resistor R2, this voltage rises to the threshold value of the CMOS inverter gate.

The gate's output then falls toward zero volts and drives transistor Q1 and relay RY1 on. The relay then remains on until power is removed from the circuit. When that occurs, capacitor C1 discharges rapidly through diode D1 and resistor R1, completing the sequence.

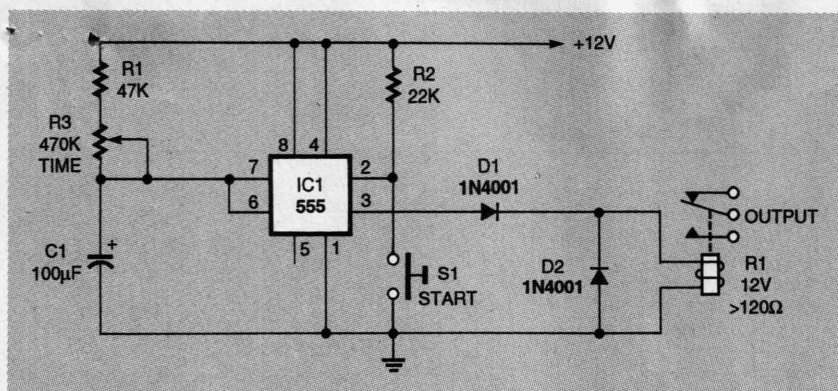


FIG. 13—SIMPLE 6-TO-60-SECOND timer relay switching circuit based on the 555 timer IC.

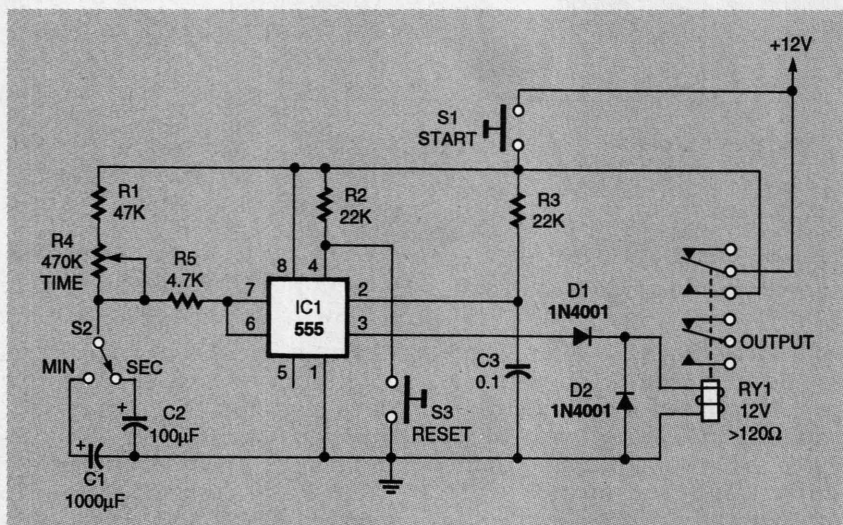


FIG. 14—TWO-RANGE 6-TO-60-SECOND and 1-TO-10 minute timer-relay switching circuit based on the 555 timer IC.

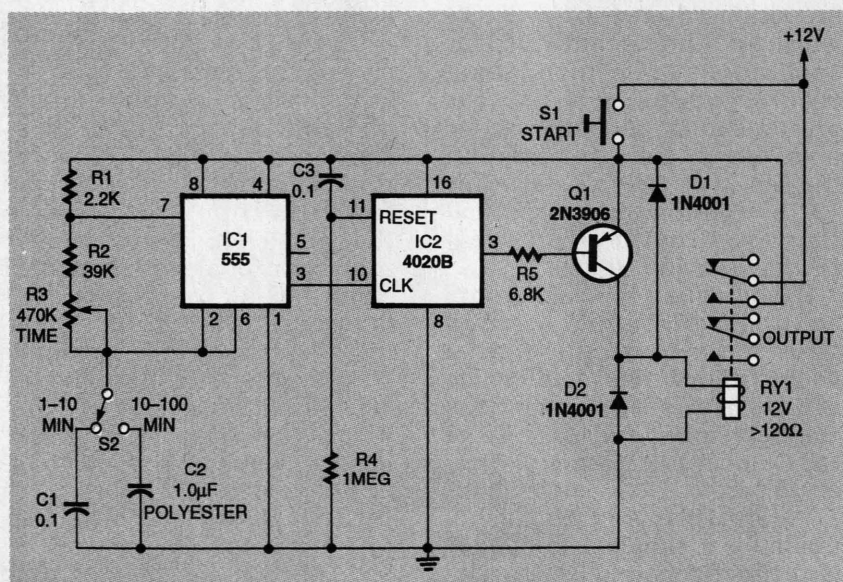


FIG. 15—TWO-RANGE 1-TO-10 MINUTE and 10-TO-100-MINUTE timer-relay switching circuit based on the 555 and 4020B ICs.

Figure 11 shows how the circuit function of Fig. 10 can be reversed so that the relay turns

on when power is applied but turns off again automatically after a preset delay. This re-

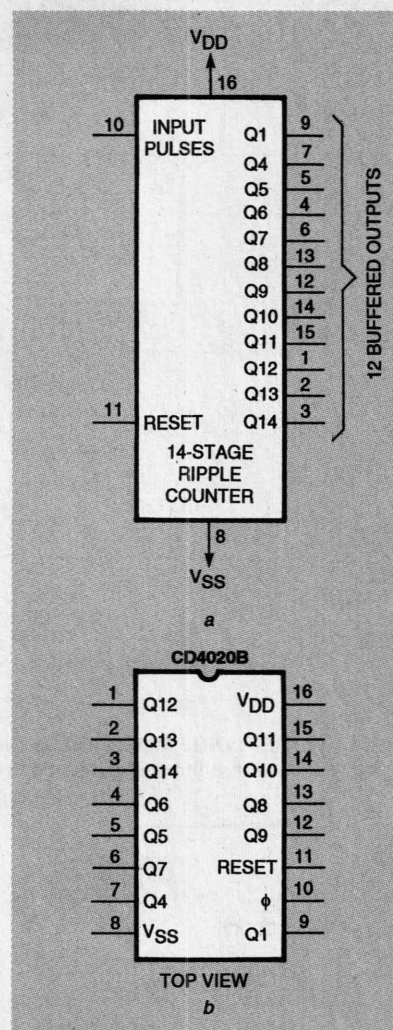


FIG. 16—CMOS 14-STAGE RIPPLE-CARRY binary counter CD4020B: functional diagram (a), and pinout diagram (b).

sponse is obtained by modifying the relay-driving stage for an NPN transistor.

It is worth noting that the circuits in Figs. 10 and 11 each provide a time delay of about 0.5 seconds for every microfarad in the value of capacitor C1. This permits delays of up to several minutes. If desired, the delay periods can be made variable by replacing resistor R2 with a fixed and a variable resistor in series whose nominal values are approximately equal to that of resistor R2.

Figure 12 shows how a pair of CMOS gates can form a push-button-activated one-shot multivibrator relay-switching circuit that provides delays up to several minutes with reasonable accuracy. The relay turns on as soon as START switch S1 is closed. However, it turns off

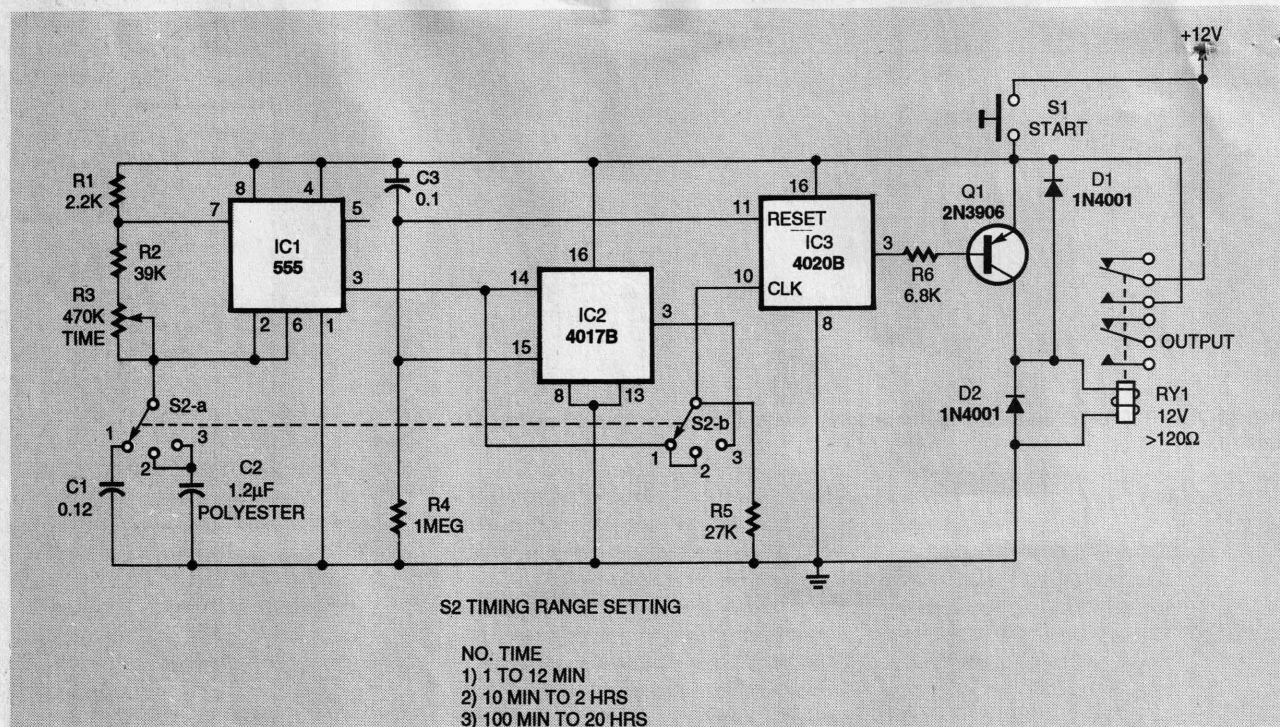


FIG. 17—WIDE-RANGE TIMER-RELAY switching circuit spans 1 minute to 20 hours in three ranges with a three-step, two-deck rotary switch.

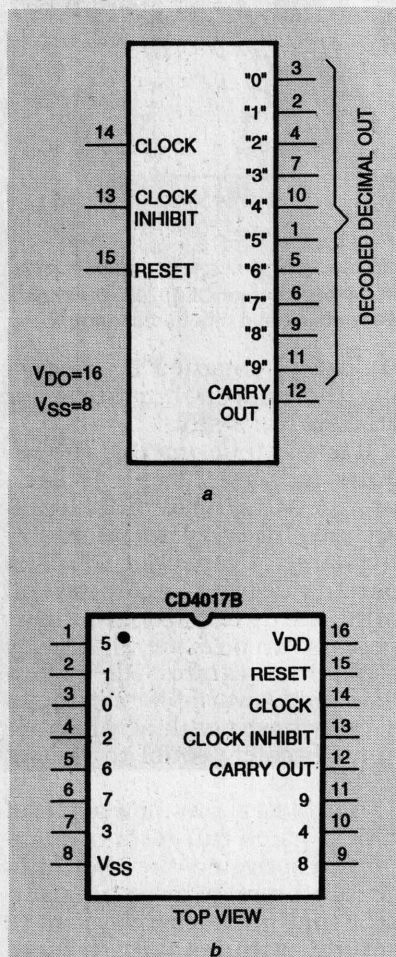


FIG. 18—CMOS FIVE-STAGE DECADE COUNTER CD4017B: functional diagram (a), and pinout diagram (b).

again automatically after a pre-set delay of about 0.5 seconds per microfarad of the value of capacitor C1. The two CMOS gates are configured as a manually-triggered monostable multivibrator whose output is fed to the relay through R4 and transistor Q1.

The circuits in Figs. 10 to 12 are all based on simple CMOS gates, and they are intended for applications where high timing accuracy is not required. Far greater timing accuracy can be obtained in circuits based on the 555-type timer IC. Figures 13 to 16 show four practical timer relay control circuits that include the 555 timer.

If you wish to learn more about how to apply the 555 timer IC (or brush up on what you do know), refer to previous articles in this series: page 58 of the September and page 69 of the October 1992 issues of *Electronics Now*.

Figure 13 is a circuit schematic for a simple 6 to 60 second timer-control circuit. The 555 IC is configured as a monostable multivibrator or a one-shot multivibrator. The circuit starts a timing cycle when START switch S1 is closed; relay RY1 is turned on immediately, and

electrolytic capacitor C1 starts to charge toward the positive power supply through 47-kilohm resistor R1 and 470-kilohm trimmer potentiometer R3.

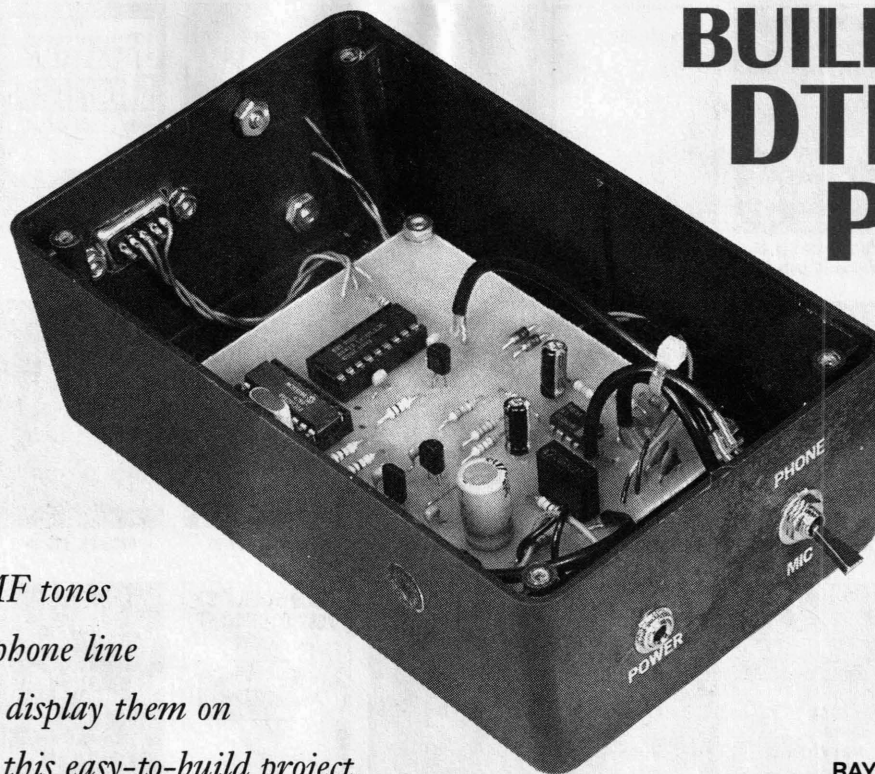
The capacitor will continue to charge until, after a delay determined by the trimmer setting, C1 rises to two-thirds of the supply voltage. At that time, the IC1 changes state and the relay is turned off. The timing cycle is then complete.

A shortcoming of the simple one-shot multivibrator relay control circuit in Fig. 13 is that it permanently draws current from the supply—even when the relay is off. Figure 14 is the schematic for a two-range timer circuit that overcomes this drawback. It is capable of timing over a range of 6 seconds to 10 minutes. The circuit operates as follows:

When START switch S1 is momentarily closed, a START pulse is sent to pin 2 of the IC1 through R3 and C3, and relay RY1 turns on. The control contacts then close, maintaining the power connections to the circuit when S1 is released. The circuit then runs through a timing cycle that is similar to the one described for Fig. 13. How-

Continued on page 90

BUILD THE DTMF- PLUS



*Decode DTMF tones
from the telephone line
or radio, and display them on
any PC with this easy-to-build project.*

RAYMOND C. BUCK, III

Decoding dual-tone modulated-frequency (DTMF) signals continues to be an area of interest for many people. DTMF signals can come from a variety of sources. They are used to control everything from home-automation equipment to amateur-radio repeaters. However, their largest use continues to be in items related to the public-telephone network.

There have been several articles published in the past that have described various ways of accomplishing the task of decoding DTMF signals. The DTMF-Plus universal decoder presented here combines the best features of those previous devices. It will be able to decode any DTMF signal from any source.

How it Works. The heart of the DTMF-Plus decoder is an SSI-202 DTMF-decoder chip. That chip determines which of the 16 possible DTMF tone-pair combinations is being received. The result is sent as an appropriate binary number to a Microchip 16C54 microprocessor. That microprocessor sends the binary information through an RS-232 port to a PC. With appropriate

software, the decoded data can be displayed on the PC's screen. There are four different data rates available that should make the unit compatible with any type of PC.

If there is a delay of more than 3 seconds between digits, the cursor on the PC is moved five spaces. If more than 10 seconds elapse between digits, a carriage return and a line feed are sent to the PC. That formatting arrangement makes the screen display more readable. For example, if someone dials a pager number, there is first a wait for the pager service to answer, after which the callback number is entered. The pager access number and the callback number will be displayed in separate columns on the screen. If more than 10 seconds elapse without dialing, the assumption is that there will be no more dialing for that particular call. If additional digits are received later, they will be displayed on a separate line. That should handle most situations.

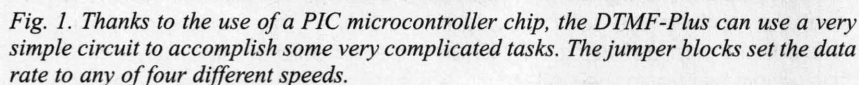
Circuit Description. There are two possible sources of DTMF tones for the circuit in Fig. 1: MIC1, an audio microphone, or J2, which connects to a standard telephone line. The inputs are individually amplified by

IC3, an LM1458 dual op-amp. That chip handles the necessary gain and impedance conversions needed by IC4, an SSI-202 DTMF detector. When the microphone is used as the signal source, some experimentation might be necessary to find the best location for placing MIC1 near a radio speaker. The gain of IC3-a is preset so that normal listening levels are sufficient for the unit to function properly.

The source of DTMF tones is routed by S1 to IC4, where conversion takes place. The binary number appearing on the output pins of IC4 is read by IC5, a 16C54 microprocessor.

The serial output of IC5 drives Q1. In order to make a stream of serial data into a true RS-232 signal, both positive and negative voltages are needed. When Q1 is off, the negative voltage at the TD terminal on the computer's serial port is connected to the RD terminal. Turning on Q1 applies positive voltage to the RD terminal. The circuitry that drives the TD terminal is protected from destruction by R15. That way, a negative-voltage is not needed.

The two jumpers labeled JP1 and JP2 set the speed of the RS-232



Power for the unit comes from a 12-volt DC wall-type adapter. The cheapest and smallest one will be fine since current requirements are very low. On-board 5-volt and 12-volt regulators IC1 and IC2 insure that the voltages needed by the individual chips are correct. As most wall-mount supplies do not have very good regulation, additional filtering for both low- and high-frequency ripple is also provided by C1 through C5.

If you decide to use perfboard, keep the wiring associated with IC3 as short as possible. With either method of construction, the connections to S1 and to the microphone should use shielded audio cable. That shielding will prevent

It is necessary for a special program to be burned into IC5 before installing it into the PC board. A pre-programmed chip is available from the source shown in the Parts List. If you want to program your own

component, the source code needed is available for downloading at the **Electronics Now** FTP site (<ftp://ftp.gernsback.com/pub/EN/dtmfplus.exe>).

The only connections to the serial port on a computer are the TD line, the RD line, and the SIG GND line. If the serial port has a 25-pin connector, connect TD to pin 2, RD to pin 3, and SIG GND to pin 7. If the serial port has a 9-pin connector, connect TD to pin 3, RD to pin 2, and SIG GND to pin 5. If you are not using an IBM-compatible computer, you will have to determine what pins to connect to by consulting your computer's manuals.

When installing the decoder in a case, it would be best to drill a hole in the case slightly smaller than the diameter of the microphone. Mounting the microphone inside the case over the hole will protect the microphone from damage if the unit is accidentally dropped. Silicone glue can be used to mount the microphone to the case.

Testing. When construction is completed, install JP1 and JP2 as needed to set the speed at which the DTMF-Plus will be communicating with your PC. Connect the DTMF-Plus to your PC and start a terminal communications program. Set up the program for the serial port to which the DTMF-Plus is connected.

The easiest way to test the unit is to connect it to a telephone line. Remove the cover from a telephone wall jack. Depending on how old the wiring is, there will be two different color code schemes in use. In the case of older wiring, the colors are green and red. With newer wiring, the colors are white wire with a blue stripe, and blue wire with a white stripe. Connect the DTMF-Plus to those wires. If you have two telephone lines in the house, line 2 might be on the white with orange stripe and orange with white stripe wires. In the case of older wiring, line 2 should be on the black and yellow wires. Polarity of the connections does not matter.

When the unit is turned on, the computer screen should display: "Touch-tone decoder is now active." If garbage characters appear on the screen, check the

PARTS LIST FOR THE DTMF-PLUS

SEMICONDUCTORS

IC1—78L05 voltage regulator, integrated circuit
IC2—78L12 voltage regulator, integrated circuit
IC3—LM1458 dual op-amp, integrated circuit
IC4—SSI-202P DTMF decoder, integrated circuit (Silicon Systems SSI75T204-IP or similar)
IC5—PIC16C54A-04P microprocessor, integrated circuit
Q1—2N4403 PNP transistor
D1—1N4001 silicon diode
D2, D3—1N5528B Zener diode

RESISTORS

(All resistors are 1/4-watt, 5% units.)
R1, R12, R13, R14—10,000 ohm
R2—R4, R10, R11—22,000-ohm
R5—33,000-ohm
R6—R9—330,000-ohm
R15—4,700-ohm

CAPACITORS

C1—220-μF, 25WVDC, aluminum electrolytic
C2—C6, C12—C14—0.1-μF, ceramic disc
C7, C11—10-μF, 16WVDC, aluminum electrolytic
C8—0.22-μF, ceramic disc
C9, C10—0.01-μF, ceramic disc

C15, C16—15-pF, ceramic disc

ADDITIONAL PARTS AND MATERIALS

J1—Wall-adaptor compatible power-input jack
J2—Modular telephone jack
J3—DB-9 or DB-25 RS-232 connector
JP1, JP2—2-position header pin
MIC1—Electret-condenser microphone cartridge
S1—Single-pole, double-throw toggle switch
XTAL1—3.579-Mhz crystal jumper blocks, 12-volt DC wall adapter, case, hardware, etc.

Note: The following items are available from ATC Electronics, PO Box 14091, Scottsdale, AZ 85260-4091: Preprogrammed PIC16C54 microcontroller, \$10.00; Printed circuit board, \$6.50; IC4 (SSI-202P), \$2.25; Complete kit of all parts, less case and RS-232 connector, \$32.50; Data-logging program, \$7.50. Please add \$4.00 for shipping and handling on all component orders—data-logging program price includes shipping. AZ residents must add appropriate sales tax. IC4 is also available from B. G. Micro, Inc., PO Box 280298, Dallas, TX 75228, Tel: (800) 276-2206.

communication program settings to make sure they are set properly. If there is no characters at all, check the wiring between the unit and the PC. Also, check the unit to make sure there are no wiring errors. If you see the sign-on message, check to make sure that S1 is in the telephone position. Pick up a telephone and dial three or four digits. Pause for 4 seconds and dial a few more digits. The two groups of numbers should appear in two separate columns. Ten seconds after the last digit is dialed, the cursor should drop down to the beginning of the next line.

Once the telephone connection is working properly, the microphone is tested next. Turn on a shortwave or amateur receiver and tune to a signal where DTMF signals can be heard. Turn the receiver volume level up to a comfortable level and place the DTMF-Plus about 6 inches from the speaker. Set S1 to the microphone position.

As DTMF signals are heard on the receiver, the corresponding numbers should be displayed on the PC screen. If a suitable receiver is not available, you could also use a speakerphone to test the microphone. Place the unit in front of the speakerphone and turn the speakerphone on. Go to another telephone, pick it up, and dial several digits. The digits should be heard over the speakerphone. The DTMF-Plus should decode them and send them to the PC. Note, however, that some speakerphones limit and clip loud signals, which could result in distortion of the tones. If that happens, the unit will not decode them, and this technique will not work.

If everything is working well, the unit is ready to be put to use. You may leave it connected permanently to the telephone line if you wish. It will not interfere with normal telephone operation and the incoming ring voltage from the

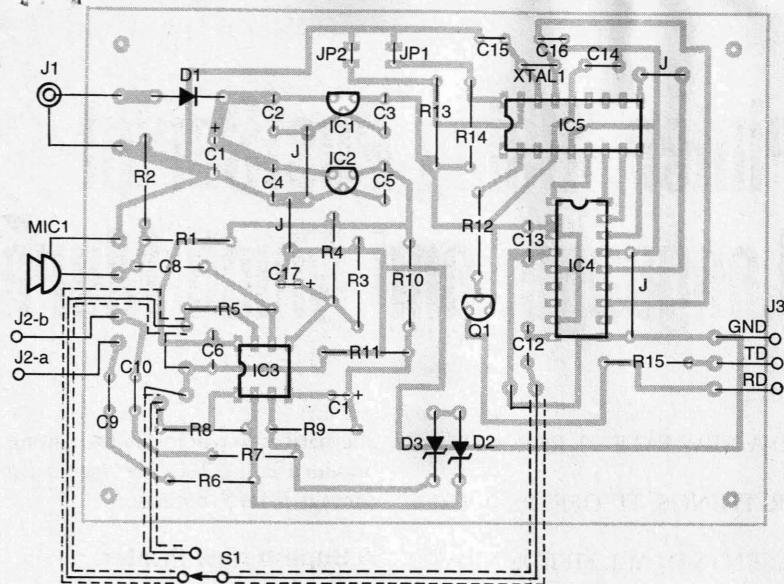
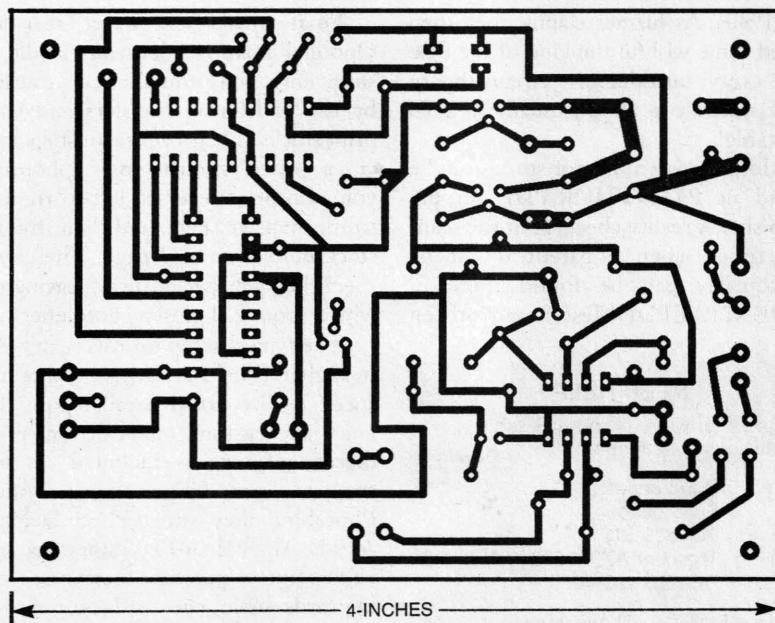


Fig. 2. Use this parts-placement diagram to locate the parts of the DTMF-Plus if you use the foil pattern. Shielded cable should be used to connect S1. The shield should be connected to the ground holes in the PC board only—do not connect them together at the switch.

telephone system will not have any adverse affect on the unit.

Using the DTMF-Plus. Data from the DTMF-Plus can be displayed on a PC using any available communications program. An optional data-logging program is also available from the source given in the Parts

List to permanently store any information that is decoded. That program lets you store any data that is decoded for viewing at a more convenient time. For example, you can leave the DTMF-Plus connected to a telephone line and a PC. Each time a number is dialed, it will be stored in the computer for later



Here's the foil pattern for the DTMF-Plus. A simple circuit and a few jumper wires easily fit the entire project onto a single-sided board.

viewing. Each time the DTMF-Plus sends a carriage return/line feed combination, the line of digits is stored as a call record in a database. Records can be deleted from the database on an individual basis or all records can be deleted at once. That DOS-based program runs on any IBM-compatible computer.

As you explore the abilities of the DTMF-Plus, you'll find new ways to use the unit. One use is as a tester for telephones, modems, and communication programs. The only limit is your imagination. Ω

SMARTBOX

(continued from page 32)

just comes on. Then accelerate slightly and note that LED2 goes out. Allow the vehicle to coast, and note that LED2 comes on again.

You will probably wish to drive several miles at various speeds and through upgrades and downgrades to learn how the setting of R14 suits your vehicle and driving habits. It may be readjusted in accordance with individual taste. Remember, when LED2 is on, the compressor is enabled when the air-conditioning system calls for cooling.

Once the calibration is completed to your satisfaction, install the Smartbox in the vehicle where the driver can operate S1 and view the LEDs.

When the vehicle is parked in full sun on a hot day, you may wish to use the override switch to obtain maximum cooling power until the interior is cool. That allows the system to operate normally without the fuel-saving feature. Once the interior of the vehicle is comfortable, switch back to automatic control for improved fuel economy and vehicle performance. Ω

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CIRCUITS THAT ARE CAPABLE OF generating a variety of waveforms are important for test and analysis in electronics. The sinewave, the most useful waveform, can be generated by various resistive-capacitive (RC) or inductive capacitive (LC) oscillators. They can also be synthesised by integrated circuit waveform generators. This article focuses on two popular, low-frequency sinewave and square wave generators, the Wien-bridge and twin-T oscillators.

Oscillator fundamentals

An oscillator, as illustrated in block diagram Fig. 1, is basically an amplifier with positive feedback. The gain equation for an amplifier with positive feedback is:

$$A_F = A/(1 + A\beta)$$

Where:

A_F = gain with feedback

A = open-loop gain

β = feedback factor, V_i/V_{sbo}

Feedback is provided by the feedback or phase-shift network, as shown in Fig. 1. The output signal is fed back to the amplifier's input terminal through a phase-shift network, where it undergoes a 180° phase shift. The amplifier also causes a 180° phase shift. The two conditions for sustained oscillation are known as the *Barkhausen criteria*.

1. The phase shift in the circuit is zero.
2. The closed-loop gain, the product of $A\beta$, must be equal to or greater than one.

An input signal is not neces-

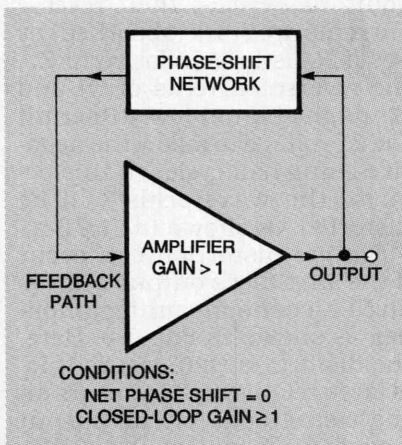
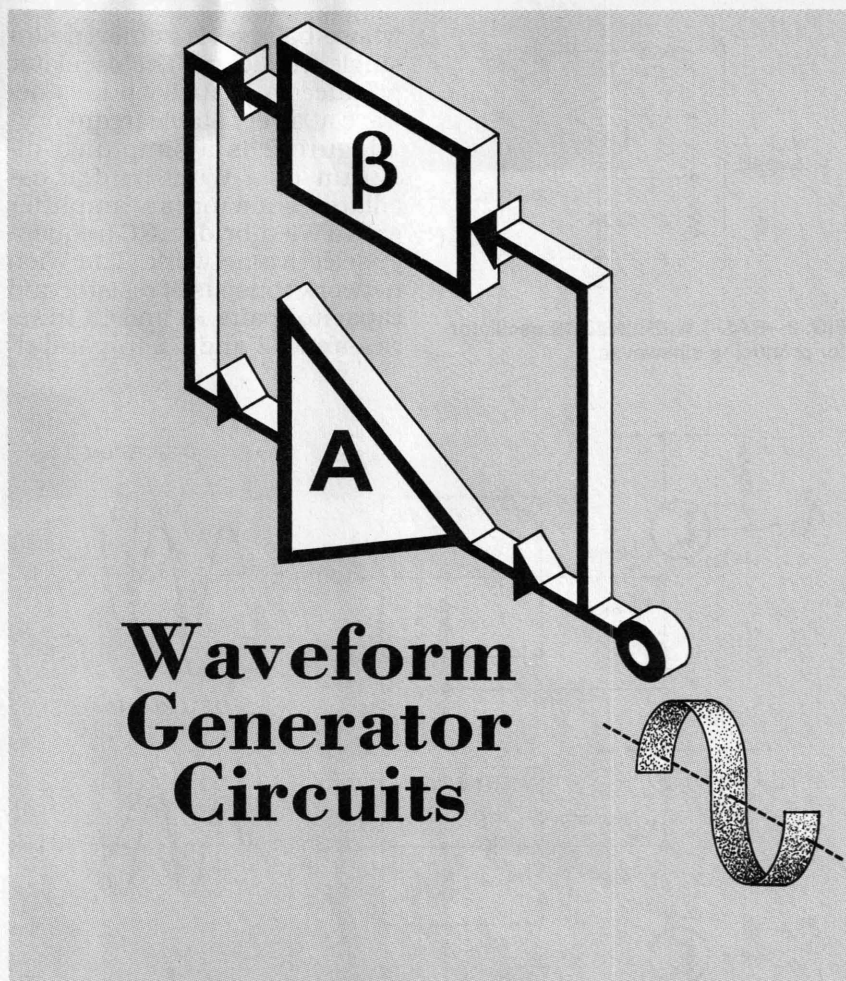


FIG. 1—CONDITIONS for oscillation based on an amplifier.



Learn about Wien-bridge and Twin-T oscillator circuits and put them to work in your circuit designs.

RAY MARSTON

sary to start oscillation—a transient is sufficient. The closed-loop gain should be greater than unity to ensure strong oscillations. However, if the gain is too high, the circuit will operate in a saturation condition, and the sinewave will be distorted.

An oscillator's frequency-selective, phase-shift network consists of a combination of resistive and capacitive components (RC) or inductive and capacitive components (LC). A combination of inductive and capacitive components is called a *tuned circuit* or a *tank circuit*.

RC oscillators are preferred to LC oscillators at low frequencies because the size and weight of

inductors at low frequencies makes the circuit too large and heavy. RC sinewave oscillators work best at frequencies from 10 Hz to about 150 kHz. This band includes the audio-frequency range of 20 Hz to 20 kHz. By contrast, LC networks work best at frequencies above 50 kHz. There is some overlap between these frequency ranges. The most common output impedances of audio oscillators are 75 ohms and 600 ohms.

Two popular low-frequency oscillators with RC phase-shift networks commonly used today are the Wien-bridge and twin-T oscillators.

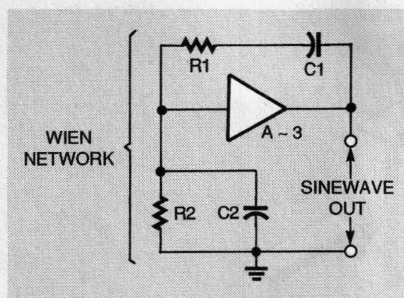


FIG. 2—BASIC WIEN-BRIDGE oscillator for producing sinewaves.

whose balance is achieved at a single frequency. The oscillator produces a relatively pure sine-wave at a very stable frequency.

Figure 2 is a simplified diagram of a Wien-bridge oscillator showing an amplifier with a Wien-bridge, RC frequency-selective network. The Wien network consists of resistor and capacitor pairs R_1 and C_1 in series and R_2 and C_2 in parallel.

precisely zero at the center frequency $f_0 = 1/(2\pi \times RC)$.

At this frequency, the network has a voltage gain of 0.33. The Wien network in Fig. 2 is connected between the output and input of a noninverting amplifier with a voltage gain of 3. Consequently, the circuit has a zero overall phase shift and a unity loop gain at the center frequency. These satisfy the Barkhausen criteria for oscillation.

A practical Wien network is unlikely to be precisely symmetrical because of the tolerances of its components, so its gain might deviate widely from the ideal 0.33 value. To compensate for this variation and permit the oscillator's loop gain to be set at unity, the amplifier's gain must be variable between about 3 and 5.

Moreover, the loop gain must initially be slightly greater than one to initiate oscillation. Then it must be reduced, either manually or automatically, to a value of one necessary to produce low-distortion sinewaves.

If the amplifier's output stage is a simple emitter follower, it might not be a low-distortion drive for the Wien-bridge network. This can be seen in Fig. 3-a, a schematic for a sinewave-driven NPN transistor emitter-follower Q_1 driving the input of a Wien-bridge network. The network is represented by capacitor C_1 and complex load Z_L .

On positive-going half-cycles, the forward currents of C_1 and Z_L are sourced (supplied) by transistor Q_1 , but on negative-going half-cycles, their reverse currents are sunk (absorbed) by R_L . If R_L is large relative to Z_L , the reverse currents of C_1 and Z_L might be too low to permit the Z_L voltage to follow the negative-going half cycles. If this occurs, the waveforms will be distorted, as shown in Fig. 3-a.

Similar distortion can occur if the amplifier's output stage is an NPN common-emitter amplifier, as shown in Fig. 3-b. Here, the distortion that occurs if R_L is large relative to Z_L appears on the rising edge of the output waveform because the C_1 - Z_L source currents flow through

Continued on page 82

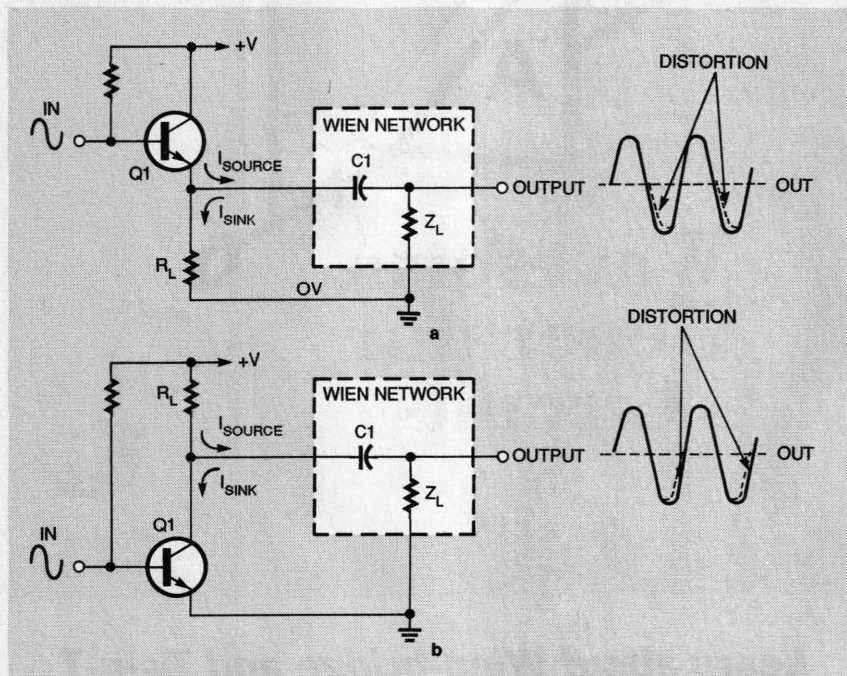


FIG. 3—WAVEFORM DISTORTION from an emitter-follower amplifier (a), and common-emitter amplifier (b), if the value of the load resistor exceeds the load impedance.

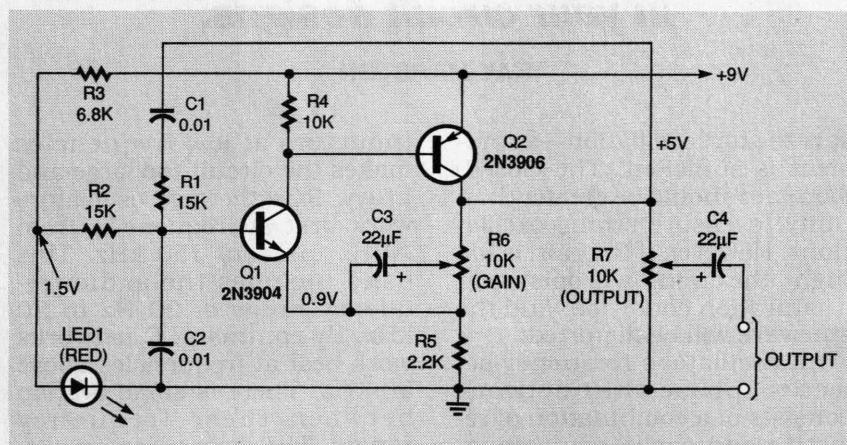


FIG. 4—THIS WIEN-BRIDGE OSCILLATOR generates variable-amplitude, 1-kHz sine-waves.

Wien-bridge oscillator

The Wien-bridge oscillator is essentially a feedback amplifier that includes a Wien bridge as its phase-shift network. The Wien bridge is an AC bridge

Here $C_1 = C_2 = C$, and $R_1 = R_2 = R$. The main characteristic of the Wien network is that the phase relation between its output and input signals can vary from -90° to $+90^\circ$, and it is

WAVE GENERATING

continued from page 36

R_L , and the sinking current flows through transistor Q1. These distortion problems do not occur with a complementary emitter follower because that circuit can source or sink high output currents with equal ease.

Transistor oscillators

Figure 4 is a schematic for a Wien-bridge oscillator based on two transistors. It will generate 1-kHz sinewaves. The oscillator draws 1.8 milliamperes from a 9-volt supply, and has an output amplitude that can be widely varied with output potentiometer R7 up to 6 volts peak-to-peak. The circuit operates as follows:

Transistors Q1 and Q2 are a pair of direct-coupled complementary feedback, common-emitter amplifiers. They present a very high input impedance to the base of Q1 and a low output impedance from Q2's collector. They also present a noninverted DC-voltage gain of 5.5, obtained from the resistor relationship of $R6 + R5/R5$. (Substituting the resistor values shown in Fig. 4 of $10K + 2.2K/2.2K$ ohms gives a gain of 5.5.)

Transistors Q1 and Q2 also provide an AC voltage gain that is variable from 1.0 to 5.5 with

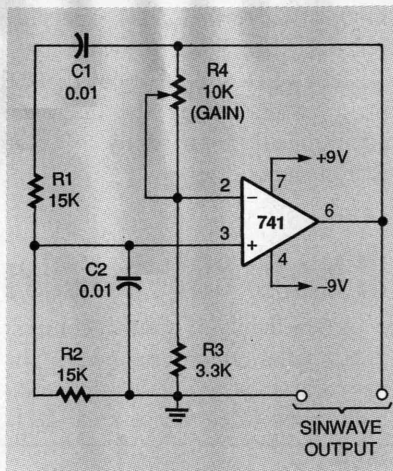


FIG. 6—A WIEN-BRIDGE oscillator based on an operational amplifier.

potentiometer R6. Transistor Q2's collector load impedance (determined by R5, R6, and R7 is approximately 5 kilohms.

Red LED lamp LED1 conducts through R3 to generate a stable, low-impedance 1.5-volt bias source. That bias is fed to the base of transistor Q1 through R2. This biases Q2's output to a quiescent value of +5 volts. The Wien network (formed by R1-C1 and R2-C2 and connected between Q2's output and Q1's input) has an active impedance of about 15 kilohms and is driven by Q2's output. Potentiometer R7 can vary the oscillator's output amplitude widely.

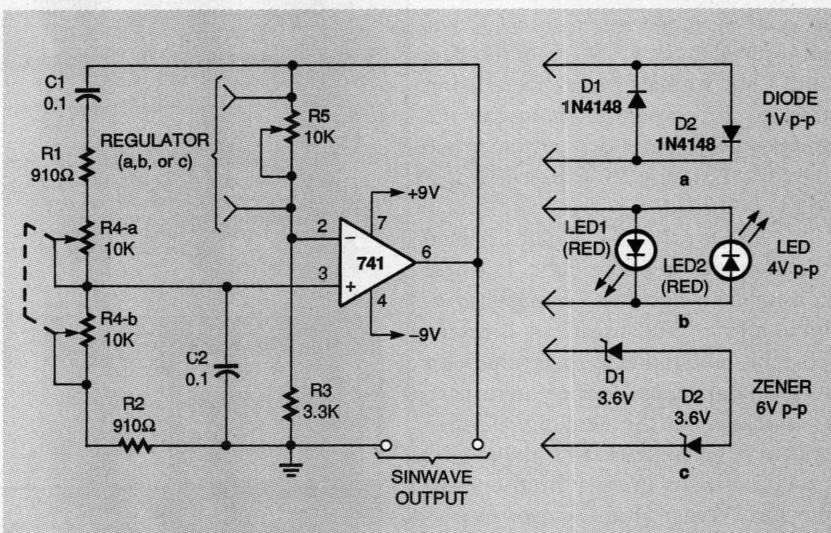


FIG. 7—AMPLITUDE OF SINEWAVES can be varied by a choice of three different voltage regulators in the 150-Hz to 1.5-kHz frequency range Wien-bridge oscillator.

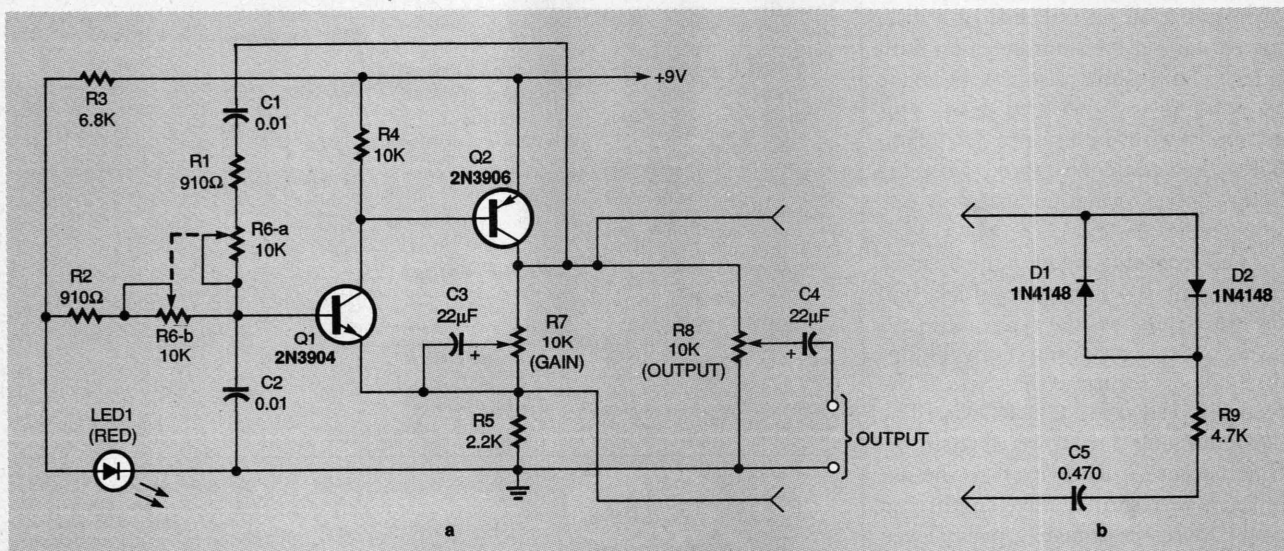


FIG. 5—THIS WIEN-BRIDGE OSCILLATOR will demonstrate its ability to generate variable frequencies between 1.5 and 15 kHz.

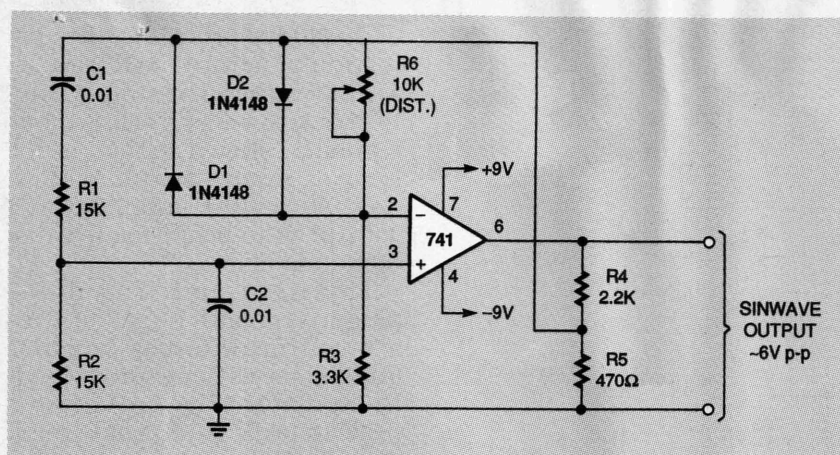


FIG. 8—A 1-kHz WIEN-BRIDGE OSCILLATOR with an amplified-diode regulator.

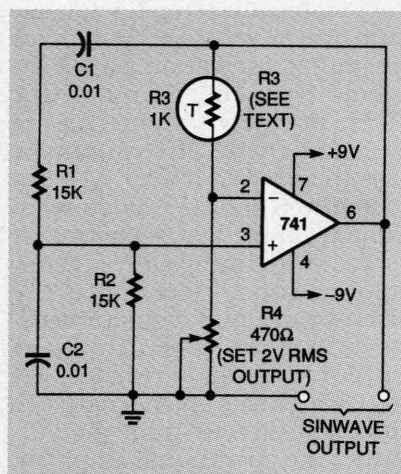


FIG. 9—A 1-kHz WIEN-BRIDGE OSCILLATOR with an NTC thermistor regulator.

To organize the oscillator in the Fig. 4 circuit, connect its output to an oscilloscope and adjust potentiometer R6 so that a stable, clean waveform is generated on the oscilloscope's screen. In this condition, oscillation amplitude is limited only by the onset of positive-peak clipping as the amplifier encounters saturation.

This clipping occurs at a peak-to-peak amplitude of about 6 volts. The clipping is a kind of distortion-generated automatic gain control (AGC) that reduces the oscillator's loop gain below the critical unity level as the waveform's amplitude reaches a specific peak value. If potentiometer R6 is carefully adjusted, this clipping can be reduced to an almost imperceptible level. The result will be high-quality sinewaves with less than 0.5% total harmonic distortion (THD).

The Wien-bridge oscillator circuit in Fig. 4 is versatile and inexpensive in fixed- and switched-frequency applications where a single-ended power supply is required. However, the values of resistors R1 and R2 must not be less than 5 kilohms. The R1-R2 and C1-C2 values can be varied over a wide range to give alternative operating frequencies. Moreover, small frequency changes can be obtained by altering the value of either R1 or R2.

Figure 5 is a schematic for a Wien-bridge oscillator, a modification of the Fig. 4 schematic. The oscillation frequency of the circuit is fully variable from 1.5 kHz to 15 kHz with dual potentiometer R6. The circuit can be both a practical sinewave generator and a Wien-bridge oscillator demonstrator. The Wien bridge is made up of C1-R1-R6-a and C2-R2-R6-b.

To demonstrate the Wien-bridge oscillator with this circuit, build it as shown in Fig. 5-a, and connect its output to an oscilloscope. Set output potentiometer R8 for maximum output, and set dual potentiometer R6 to its maximum resistance value to provide the lowest frequency. Then carefully adjust potentiometer R7 so that clean, stable waveforms appear on the oscilloscope. When set up this way, the Wien network presents a high impedance and is easily driven by transistor Q2. As a result, oscillation amplitude is limited only by the onset of positive-peak clipping at an amplitude of about 6 volts, peak-to-

peak, as described earlier.

Progressively increase the operating frequency by reducing the resistance of R6 while retrimming R7 as necessary to maintain oscillation, until the maximum frequency is obtained. Under this condition, the waveform's peak-to-peak amplitude drops to about 3 volts, and slight distortion is visible on the falling halves of the waveform. This is caused by transistor Q2 losing its ability to drive the Wien network cleanly as its impedance drops below a few kilohms. The result is a form of amplitude-limiting, distortion-generated AGC.

This distortion can be overcome, at the expense of a large increase in the circuit's quiescent current, by inserting a 1.5-kilohm resistor between transistor Q2's collector and ground. This increases Q2's output drive capability. However, with this modification, the waveform's amplitude is again limited by clipping at about 6 volts, peak-to-peak.

To finish setting up the network as a demonstration unit, remove the 1.5-kilohm resistor from Q2's collector, and connect the network formed by diodes D1 and D2, resistor R6 and capacitor C5, as shown in Fig. 5-b across potentiometer R7, at the two jacks.

Vary the wiper of potentiometer R6 across the whole frequency band, and trim potentiometer R7 to find a position where you obtain a clean sinewave with an amplitude of about 1.2 volts peak-to-peak, at all settings of R6. Resistor R6 and the two diodes, D1 and D2, introduce a form of gain control. At the start of each half-cycle, the circuit's loop gain is greater than one, so oscillation starts. However, as soon as the waveform amplitude approaches a peak value of about 600 millivolts, either D1 or D2 will conduct, shunting 4.7-kilohm resistor R9 across R7, thus reducing the loop gain to unity.

This is a gentle form of limiting that causes less than 1% THD. The version of Fig. 5 circuit with the D1, D2, R9, and C5 network plugged in becomes an

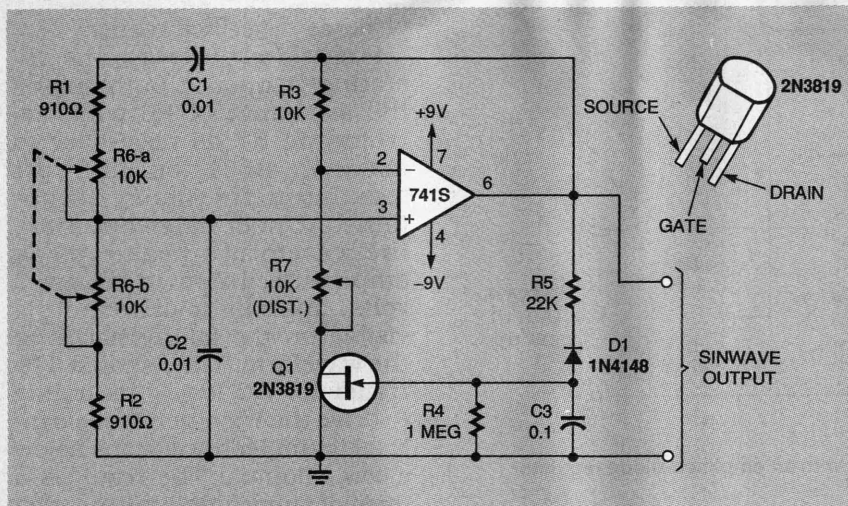


FIG. 10—A JFET-REGULATED 1.5- to 15-kHz Wien-bridge oscillator.

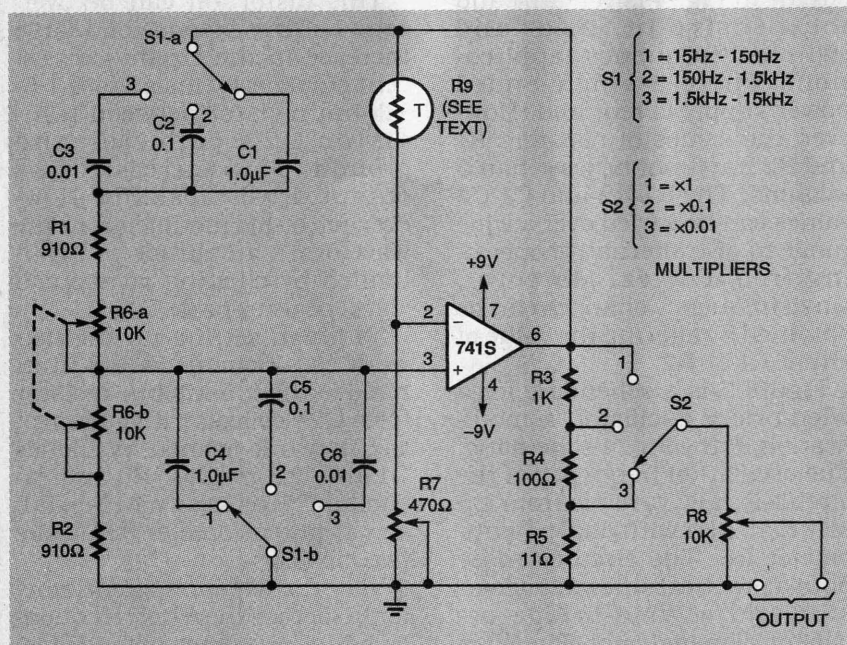


FIG. 11—A THREE-DECADE (15 Hz to 15 kHz) Wien-bridge oscillator.

inexpensive but effective variable-frequency sinewave generator whose frequency range can be changed by substituting different values for capacitor C1 and C2.

Oscillators with op-amps

Standard integrated-circuit operational amplifiers offer high voltage gain, high input impedance, and low output impedance. They are excellent amplifiers for Wien-bridge oscillators. Figure 6 is a schematic for a simple 1-kHz oscillator. Resistor-capacitor pairs R1-C1 and R2-C2 form the Wien-bridge network. Trimmer potentiometer R4 and resistor R3 control

the operational amplifier's closed-loop gain.

The circuit oscillates when R4 is adjusted, but its waveform amplitude is limited only by the onset of peak clipping as the op-amp encounters saturation. If R4 is adjusted carefully, clipping can be reduced to an insignificant level, so that the output sinewaves are clean. However, this technique is useful only in fixed-frequency applications. Where variable frequency is desired, some form of gain control must be provided so that there will be amplitude limiting with minimal distortion.

Figure 7 shows three ways to regulate waveform amplitude

with diodes providing the distortion-generated AGC, as described earlier. The simple plug-in, back-to-back 1N4148 diode regulator, shown in 7-b, is the least expensive solution. However, its output is limited to 1 volt, peak-to-peak, for low-distortion levels.

The light-emitting diode plug-in regulator (Fig. 7-c) gives a 4-volt peak-to-peak output and the series Zener diode plug-in regulator (Fig. 7-d) gives a peak to peak output of 6 volts. With any of these plug-in regulators in position, adjust trimmer potentiometer R5 for the minimum setting that gives sustained oscillation across the whole frequency band. If the Wien bridge components are well matched, THD might be less than 0.5%.

Figure 8 shows a diode-regulated Wien-bridge oscillator that is well suited for fixed-frequency applications. Here, both the Wien-bridge and regulator-feedback loop are taken from the output junction of resistors R4 and R5. This is an alternative to taking it directly from the output of the operational amplifier. This simple modification effectively amplifies the diode regulation voltage by a factor determined by the ratio of the value of $(R4 + R5)/R5$. In the Fig. 8 circuit, the sinewave output is about 6 volts, peak-to-peak. If trimmer R6 is adjusted carefully, THD levels of 0.1% can be obtained.

Linear AGC circuits

The Wien-bridge oscillators shown in Figs. 6, 7, and 8 rely on distortion-generated AGC to control waveform amplitude. Those circuits can maintain a constant and jitter-free amplitude as the frequency is increased and decreased across the available band.

There is an alternative based on automatic linear gain control. Oscillators with this feature usually generate negligible distortion. However, they suffer from amplitude "bounce" when the frequency settings are shifted up and down the available band. The AGC system hunts for the correct gain value. Fig-

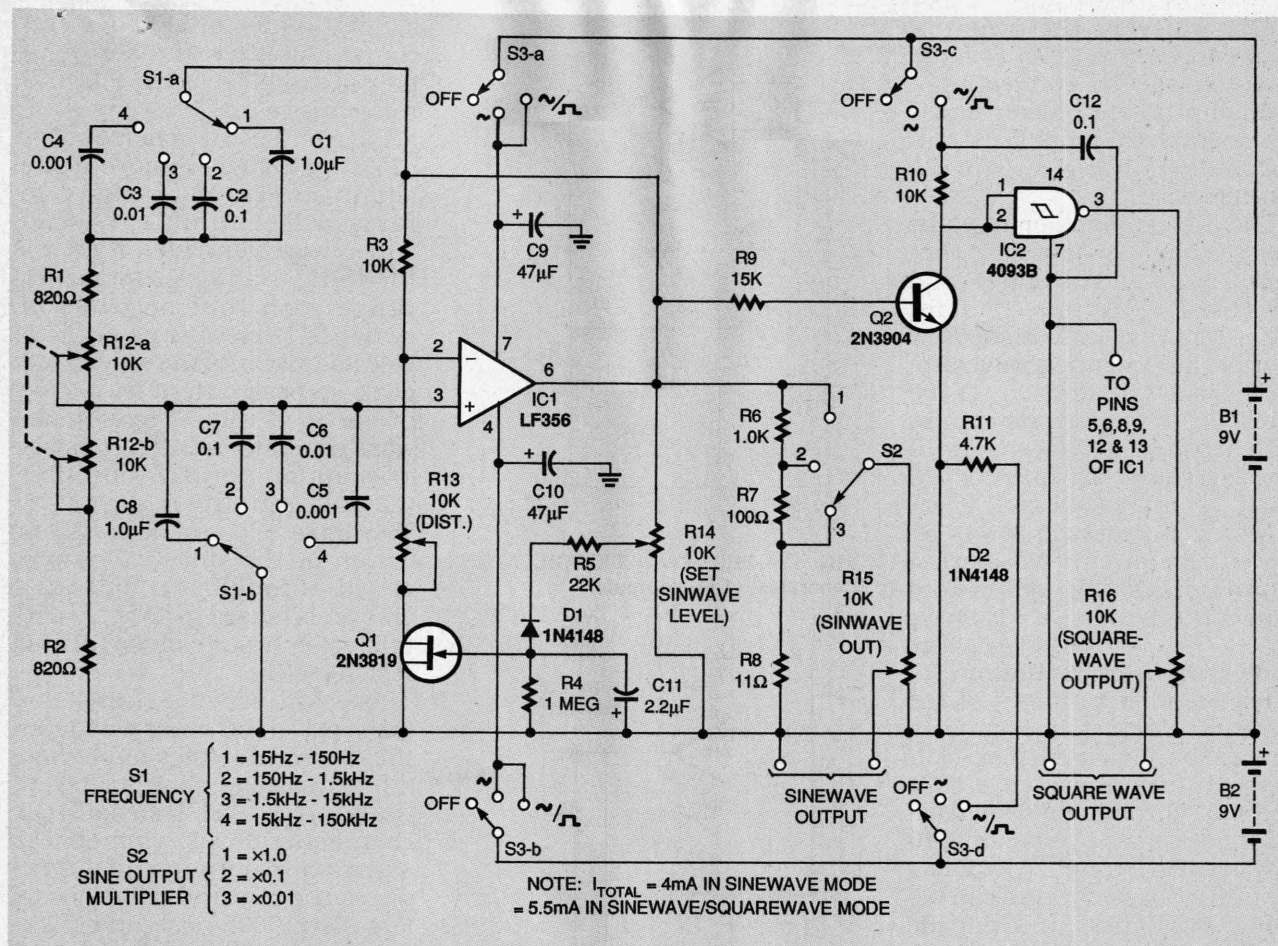


FIG. 12—THIS SINE/SQUAREWAVE generator covers the 15-Hz to 150-kHz band.

ures 9, 10, and 11 show three circuit schematics that contain linear AGC circuits.

Figure 9 is the schematic for a 1-kHz oscillator circuit whose output amplitude is stabilized by a negative temperature coefficient (NTC) thermistor. It is shown here as R3, a resistor symbol in a circle containing the letter T. The resistance of an NTC thermistor decreases as temperature increases. Typical resistance values for NTC thermistors devices range from 200 ohms to 50 kilohms. Thermistor R3 in series with trimmer potentiometer R4 becomes a gain-determining feedback network. The value of R3 is 1 kilohm.

Thermistor R3 is heated by the output signal's average power. At the desired amplitude, R3 has a resistance value double that of the 470-ohm trimmer potentiometer R4, thus giving the circuit an overall gain of unity. If the output amplitude starts to rise, R3's

temperature rises and its resistance falls. This response reduces the gain and restores the original output level. The reverse action occurs if the output starts to fall, and the original output level is again restored. This oscillator generates negligible distortion.

Figure 10 is the schematic for an alternative circuit that consumes only a few milliamperes. In this circuit, JFET Q1 serves as a variable resistance that is voltage-controlled by the oscillator's negative peak amplitude. This is detected by the series network of resistor R5, diode D1, and capacitor C3 in parallel with resistor R4.

The drain-to-source path of JFET Q1 acts as a low resistance when Q1's gate is biased to zero volts. However, the resistance of the path approaches infinity when Q1's gate is biased to a negative pinchoff value of a few volts. When dial potentiometer R6 is suitably adjusted, the circuit oscillates and generates

very little distortion in its output with a peak-to-peak value of X volts.

If the output tries to rise above a set value (X), the detected change automatically increases Q1's resistance value. This response reduces the operational amplifier's gain, bucking the attempted rise in output. However, if the output tries to fall below the X value, an opposite response occurs—gain increases to maintain a constant output level.

To organize the Fig. 10 circuit, connect its output to an oscilloscope and adjust trimmer R7 to the lowest setting that will give stable oscillation without visible distortion of the waveform over the 1.5 kHz to 15 kHz band. If the Wien bridge components are well matched, THD could be less than 0.1%. The output signal amplitude depends on the pinchoff characteristics of the 2N3819 JFET, a value that is typically between 2.5 and 7 volts, peak-to-peak.

Alternatively, the pinchoff level can be preset to 8 volts, peak-to-peak, by connecting a 10-kilohm potentiometer across the operational amplifier's output and supplying resistor R5 from its wiper.

The AGC system's charge/discharge time constants are controlled by R5 and C3 and R4 and C3. The R4-C3 time constant must be long with respect to that of the generated waveform cycle for low distortion. If the oscillator is to generate signals with frequencies as low as 15 Hz, the value of capacitor C3 must be increased to 2.2 μ F. However, if capacitor C3 is removed, the circuit will have a simple distortion-generated gain control.

Wide-frequency oscillator

The frequency ranges of the oscillator circuits in Figs. 6 to 10 can be changed with different values of capacitors C1 and C2. For example, increasing those values by a factor of 10 reduces the frequency by a factor of 10. Those circuits can become wideband multidecade oscillators by installing switches for selecting alternative decade capacitor values.

The maximum useful operating frequency for this kind of circuit is restricted by the slew rate limitations of the operational amplifier. The useful frequency limit for the low-cost 741 operational amplifier is 20 kHz. That limit is raised to 80 kHz with a 741S, 120 kHz with a National Semiconductor LF355, and 250 kHz for a National Semiconductor LF356.

If you build a variable-frequency Wien-bridge oscillator, be sure that the two tracks of the frequency-control potentiometer are well matched if you want low-distortion and stable-amplitude performance. In multidecade oscillators, the values of the Wien-bridge capacitors should be as closely matched as is practical on all ranges. If these capacitors are not well matched, it might be necessary to install an AGC GAIN or DISTORTION switch for each range.

Figure 11 is the schematic for

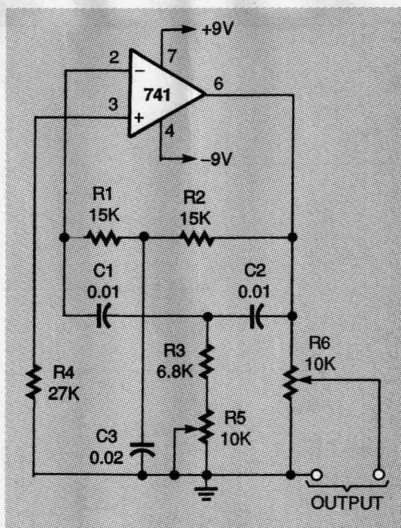


FIG. 13—THIS TWIN-T OSCILLATOR generates a 1-kHz output.

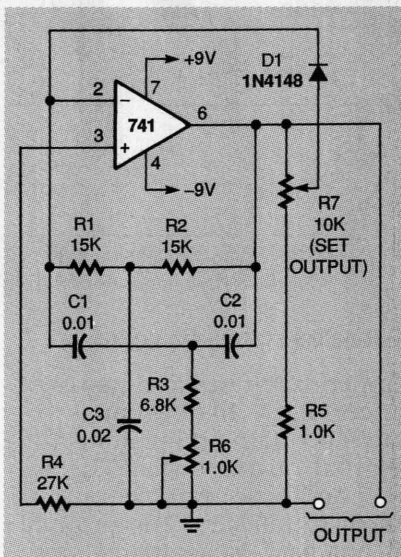


FIG. 14—A TWIN-T OSCILLATOR with diode regulation.

a variable-frequency Wien-bridge oscillator whose output covers the 15-Hz to 15-kHz frequency band in three switched decade ranges. The circuit includes thermistor stabilization, as previously described, and it generates an output with low-distortion. The frequency output of the oscillator can be switched with rotary switch S1 in three bands: 15 Hz to 150 Hz, 150 Hz to 1.5 kHz, and 1.5 kHz to 15 kHz. The amplitude of the output signal can be changed with switch S2.

The frequency limit of this oscillator can be increased to 150 kHz by replacing the 741S operational amplifier with a Nation-

al Semiconductor LF356 operational amplifier and adding a pair of switch-selected 0.001 μ F range capacitors.

Figure 12 is the schematic for JFET-regulated version of the circuit in Fig. 10, modified to form an economical sine/squarewave generator. It covers the 15 Hz to 150 kHz frequency range with four settings of switch S1. Sinewave potentiometer R14 permits the sinewave's peak-to-peak output level to be preset to 8 volts. The square-wave generator section consists of common-emitter amplifier Q2 and Schmitt trigger IC2. The base of transistor Q2 is driven by the sinewave output of IC1 through R9, and Q2's emitter is biased to about -600 millivolts through diode D2 and resistor R11.

As a consequence, transistor Q2 is driven on or off whenever the sinewave swings more than a few millivolts above or below the zero-volt level. Transistor Q2 then generates a symmetrical squarewave output, and IC2 reduces the rise and fall times to less than 100 nanoseconds. The final squarewave output is available from R16. Switch S2 is the output multiplier. To conserve battery power, the square-wave generator can be switched off with switch S3-b and S3-d when the circuit is not in use.

Twin-T oscillators

The twin-T oscillator is another popular RC network-based sinewave oscillator that is useful in fixed-frequency applications. It is made by connecting a twin-T network between the output and input of an inverting operational amplifier, as shown in Fig. 13. The twin-T network consists of resistors R1 and R2 shunted by capacitor C3, and capacitors C1 and C2 shunted by resistor R3 in series with potentiometer R5.

These components form a balanced network because they are in the ratios $R1 = R2 = 2(R3 + R5)$, and $C1 = C2 = C3/2$. They give a zero output at a center frequency, f of $1/2\pi \times R1 \times C1$, and a finite output at all other frequencies. If

the Twin-T is imperfectly balanced, it gives a feeble output at its center frequency, and its output phase depends on the direction of the imbalance. In other words, if the imbalance is caused by low values of $R3 + R5$, the output phase is inverted with respect to the input.

The twin-T network in Fig. 13 can be adjusted by $R5$ so that a small phase-inverted output at a center frequency of 1 kHz can be produced. Consequently, overall phase inversion takes place around the feedback loop, and the circuit oscillates at 1 kHz. Potentiometer $R5$ can be adjusted so that oscillation is barely sustained. Under this condition, sinewave amplitude is limited to about 5 volts RMS by the onset of operational amplifier clipping. The output has less than 1% THD.

Figure 14 is a schematic for a simple twin-T oscillator that will have lower output distortion. Diode $D1$ provides distortion-generated automatic gain control. To organize this circuit, set the wiper of potentiometer $R7$ to the output of the operational amplifier, and adjust $R6$ so that oscillation is just sustained. A sinewave output of about 500 millivolts, peak-to-peak, will be produced. Ω

JACOB'S LADDER

continued from page 59

to those found on the schematic Fig. 1. The cores and bobbins for this transformer are from Philips (Ferroxcube) Components, Saugerties, NY). The cores are No. E187 made from 3E2A ferrite material. The plastic molded bobbins are Part No. E187PCB1-8. These parts are available from the source given in the Parts List.

Transformer T1: Wind the first 30 turns trifilar (three wires in parallel) from No. 30 AWG magnet wire and the remaining 30 turns as single wire turns. Solder the ends of the windings to the pins, insert the E-core, and tape the assembly together securely.

Transformer T2: The core of transformer is No. 4162S from Samhwa USA, Chatsworth CA, with matching seven-segment bobbin, cup and primary bobbin. These parts are also available from the source given in the Parts List.

Start winding the secondary bobbin by securing the magnet wire to a pin insulated on the bobbin end and wind between 300 and 400 turns of No. 37 AWG magnetic wire on each of seven segments. Snap the cup in place and solder the output leads. Fill the cup with epoxy or RTV silicone when you are satisfied that the transformer has been wound correctly.

Wind eight turns of No. 30 twisted wires on the primary bobbin and tape the windings in place. Form the air gap of the primary section with 0.005-inch thick Milar tape and form the air gap for the secondary section with 0.010-inch thick Milar tape. Tape or clamp the two sections together with a heavy rubber band. Ω

VIDEO NEWS

Continued from page 6

Digital Camcorders Here

If you have a couple or three thousand dollars to spend on really top-notch video reproduction, you now can buy a digital camcorder using the standard DV (digital video) format from either Sony or Toshiba, with JVC waiting in the wings. In this column we've carried a lot of information about that DVC format (now called DV). The available camcorders produce breathtaking pictures and are ideal for professional or "prosumer" videographers.

Panasonic's model has three CCD pickups, while Sony has two models, one with three pickups and one with a single CCD. The Sony model has a digital interface and utilizes an optional feature of the DV system—a cassette with a built-in chip that stores data on tape specifications and records the date of recording and provides easy access to any scene. **EN**

IR LOGIC PROBE

continued from page 72

prepare the cable ends in the same manner as before. Solder the leads of the cable to the circuit board as shown in Fig. 2. Attach a 9-volt battery to the connector and turn on the power. The unit can be tested with an infrared remote control or an infrared LED and current-limiting resistor connected to the output of a pulse or function generator. (The IR LED can be coupled to the probe's tip with a length of 1/4-inch heat-shrink tubing for test purposes.)

The tricolor LED should glow red; if not check $D1$ and the battery or its polarity. Point the IR LED at the probe's tip and make sure both LED indicators are functioning. If there is no indication, check the photodiode's polarity, $IC1$, and its associated components. If the tricolor LED does not flash amber, check $IC2$ and its associated components. Connect an oscilloscope to $J1$, and verify that a waveform is present.

Once the circuit has been checked out, join both halves of the enclosure. Be sure that the battery wires do not tangle around the LED indicators. A piece of foam rubber cut to fit the bottom of the battery compartment will keep the battery from rattling around in the case.

Using the probe

If light from windows and incandescent lamps interferes with the probe's operation, either remove the source of interference or move the work area. Ambient room light should not create problems, especially if its source is fluorescent fixtures.

The probe's plastic light pipe is most sensitive at the screwdriver-shaped tip. The light reflects within the walls of the light pipe (as in fiber optic cable), conducting the light to the photodiode. The probe will indicate if high-frequency pulsed IR LEDs are working, but the photodiode will integrate the pulses to a steady-state level. Ω

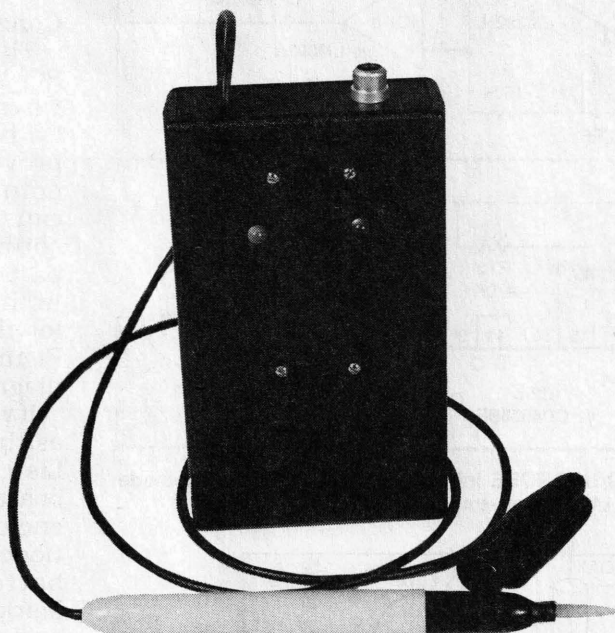
IF YOU HAVE EVER TRIED to test an infrared LED with a fluorescent phosphor card, you'll be happy to know that there's a better way. This project, called the infrared logic probe, combines an infrared photodiode sensing circuit and a logic-probe pulse detecting circuit. The device is handy for checking just about any infrared emitting source.

The infrared logic probe consists of two sections: a probe and a PC board containing the electronics. The probe is packaged in a felt-tip pen case. The electronics are packaged in a plastic case and connected to the probe through a thin coaxial cable. The circuit will detect 0.3-milliwatt continuous levels and pulses as narrow as 40 microseconds at a frequency of 7.1 kilohertz. The probe's tip is small enough to fit in a slotted optical switch and other hard-to-reach optical sensing devices. Sensitivity to ambient light is not a problem, but the probe can be sensitive to sunlight or incandescent light that is rich in infrared. The photodiode is packaged in a visible-light rejecting case with a peak spectral response of 925 nanometers and a usable range of 725 to 1150 nanometers.

Circuit description

The schematic for the IR logic probe is shown in Fig. 1. Infrared light detected by photodiode D2 is amplified by IC1-a, half of an LM392N op-amp. Resistors R1 and R2 set the voltage gain of IC1-a. The value of R2 can be changed to decrease the sensitivity of the circuit if your application demands it. Con-

INFRARED LOGIC PROBE



***Here's a device to help you
troubleshoot infrared
emitters. You can't buy it
anywhere—you have
to build it yourself!***

ALEXANDER D. FIRMANI

necting J1 provides an output to an oscilloscope for the display of the amplified photodiode signal. This is handy when checking the pulsed emitters found in most remote controls.

Voltage comparator IC1-b squares up signals from IC1-a to digital logic levels for IC2-a. Resistors R4 and R5 set the reference voltage at the non-inverting input to one half of the supply voltage, and R6 provides hysteresis to prevent oscilla-

tions. Resistor R8 pulls up the comparator's output for a near rail-to-rail voltage swing for IC2-a. LED1 and current-limiting resistor R7 indicate the presence of steady-state infrared and also function with pulsed emitters, if the duty cycle is appropriate.

Monostable multivibrator IC2-a conditions pulse trains with any period shorter than the time constant of R9 and C1 into a low-frequency waveform with a very high duty cycle. Monostable IC2-b triggers on the waveform from IC2-a. This provides pulses for LED2 that are constant in frequency and duty cycle, regardless of the high input frequency to IC2-a. Any frequency input to IC2-a with a period longer than the time constant of R9 and C1 creates IC2-b output pulses with the same width as before at the input frequency. Resistor R10 and C2 set the output pulse width for IC2-b.

Tricolor LED2 (a dual red/green device) functions as a pilot lamp and indicator for pulsed infrared sources. LED2 will always glow red and pulse amber (red + green) when infrared pulses are detected. Transistor Q1 is

an emitter-follower buffer that allows IC2-b to drive the green diode. Resistors R11 and R12 limit current for LED2.

The power source for the circuit is a 9-volt battery. Diode D1 protects the circuit from accidental voltage reversals when you install the battery. Power supply noise is decoupled by C3 and C4. Alkaline batteries will provide many hours of operation, because the circuit has low-power integrated circuits

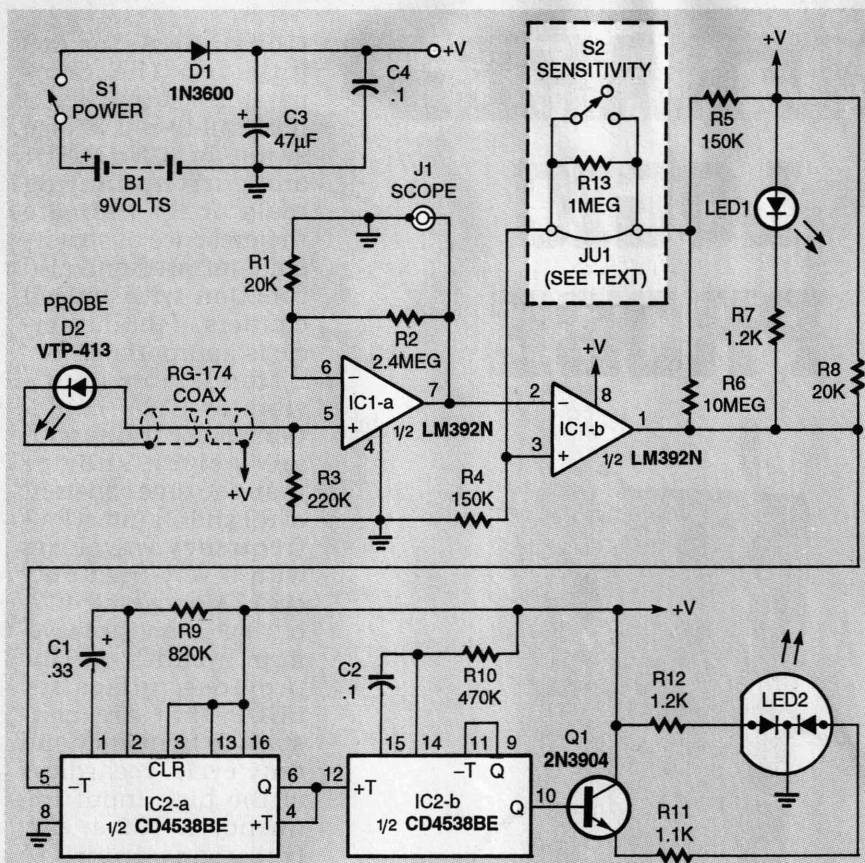


FIG. 1—SCHEMATIC FOR THE IR LOGIC PROBE. Infrared light detected by photodiode D2 is amplified by IC1-a, half of an LM392N op-amp.

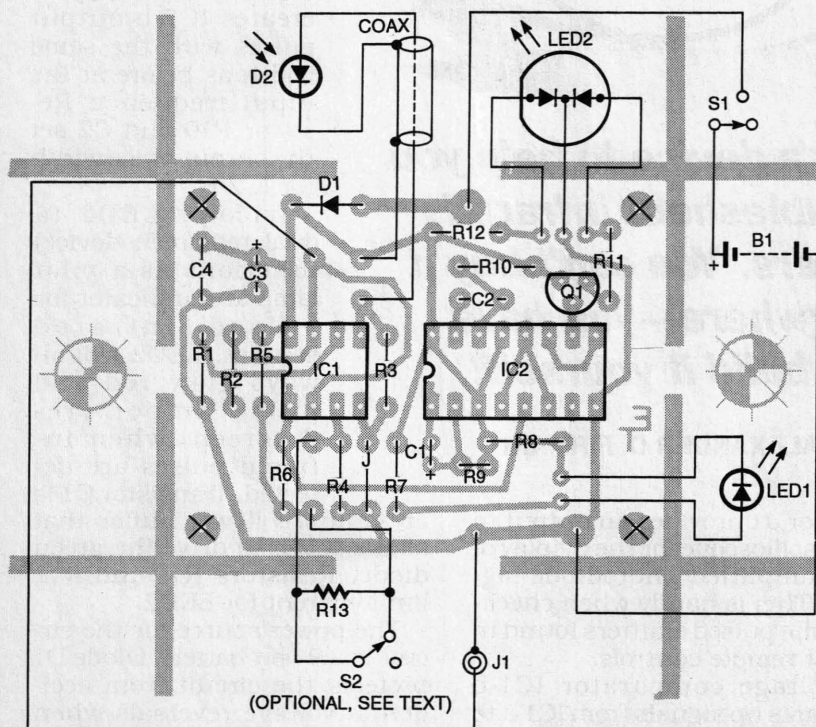


FIG. 2—PARTS-PLACEMENT DIAGRAM. Sockets for mounting the integrated circuits are recommended.

place of S2 for most emitters (remote controls) found on consumer electronic equipment. For certain devices such as slotted optical switches, CD laser diodes, and reflective sensors, more sensitivity might be desirable. If you plan to use the probe for LEDS that operate below 0.5 milliwatts, install S2 and R13—if not, you can install a wire jumper on the board instead of the switch.

Construction

The circuit can be built on a PC board or point-to-point wired. You can make your own PC board from the foil pattern provided here, but point-to-point wiring is practical because of the low component count. The photograph and parts placement diagram within this article will help. Sockets for the ICS are recommended. Figure 2 is the parts-placement diagram.

If you are using the same case as the prototype (see the Parts List), place the unpopulated PC board on the bottom half of the enclosure, centered and positioned about 5/16-inch from the battery compartment wall. Mark the four mounting holes and then drill them for 2-56 hardware.

Mount the components on the board, taking care not to make any solder bridges or poor connections. Note that R9 must be mounted vertically. Again, if you are using the recommended case, cut the leads on the LEDs to approximately $\frac{5}{8}$ -inch, and solder them to the board straight up. (If you are using a case with a different height, cut the LED leads to a length so that they will just extend through holes drilled in the top of the case after the board is mounted in the case.) Next solder 6-inch lengths of No. 24 wire for the switch/switches (remember that S2 is optional), jack J1, and the negative lead of the battery snap, to the points indicated in Fig. 2. Do not connect the switches and jack now. Mount the board to the bottom of the case with short 2-56 machine screws and nuts.

Drill two holes in the en-

and high-efficiency LEDs.

Switch S2 (and R13) is op-

tional. The probe will operate properly with a wire jumper in

PARTS LIST

All resistors are $\frac{1}{8}$ watt, 5%.

R1, R8—20,000 ohms

R2—2.4 megohms

R3—220,000 ohms

R4, R5—150,000 ohms

R6—10 megohms

R7, R12—1200 ohms

R9—820,000 ohms

R10—470,000 ohms

R11—1100 ohms

R13—1 megohm (optional, see text)

Capacitors

C1—0.33 μ F, 50 volts, electrolytic

C2—0.1 μ F, 50 volts, Mylar

C3—47 μ F, 16 volts, electrolytic

C4—0.1 μ F, 25 volts, ceramic disk

Semiconductors

IC1—LM392N dual op-amp/comparator National Semiconductor

IC2—CD4538BE dual monostable multivibrator

Q1—2N3904 NPN transistor

D1—1N3600 or NTE-519 silicon diode

D2—Vactec VTP-413 photodiode (Allied Electronics)

LED1—High-efficiency yellow LED

LED2—Tricolor LED (Digi-Key P391 or equivalent)

Other components

S1—SPST slide switch

S2—SPST slide switch (optional, see text)

J1—Chassis mount phono jack

B1—9-volt alkaline or lithium battery

Miscellaneous: One 8-pin low profile IC socket, one 16-pin low-profile IC socket, 4 feet of RG-174 coaxial cable, 9-volt battery connector, Pac Tec HM-9VB or Radio Shack 270-293 plastic project case, 2-56 hardware, clear casting resin and catalyst or clear RTV silicone sealer, 5-minute epoxy, Sanford *Sharpie* fine-point permanent marker pen (use a dried up one if you have one), small piece of $\frac{1}{16}$ -inch thick clear Plexiglas, denatured alcohol, black paint, foam rubber, No. 24 hook up wire, solder.

closure's removable end plate: one $\frac{1}{8}$ -inch hole for the probe's cable and one $\frac{1}{4}$ -inch hole for phono jack J1. Install J1 in the end plate and place the plate into the enclosure. Solder the wires to the jack, observing polarity and taking care not to melt the plastic end plate.

Next use the drill guide in Fig. 3 to mark and drill the holes in the enclosure's top. (The drill guide matches the PC board layout, so it can be used for any case.) Remember, S2 is optional, so don't drill holes for

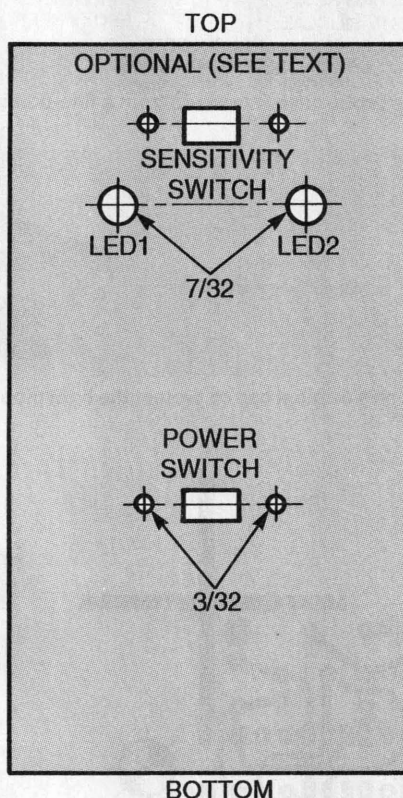


FIG. 3—DRILL GUIDE. S2 is optional, so don't drill holes for mounting it if you're not using it.

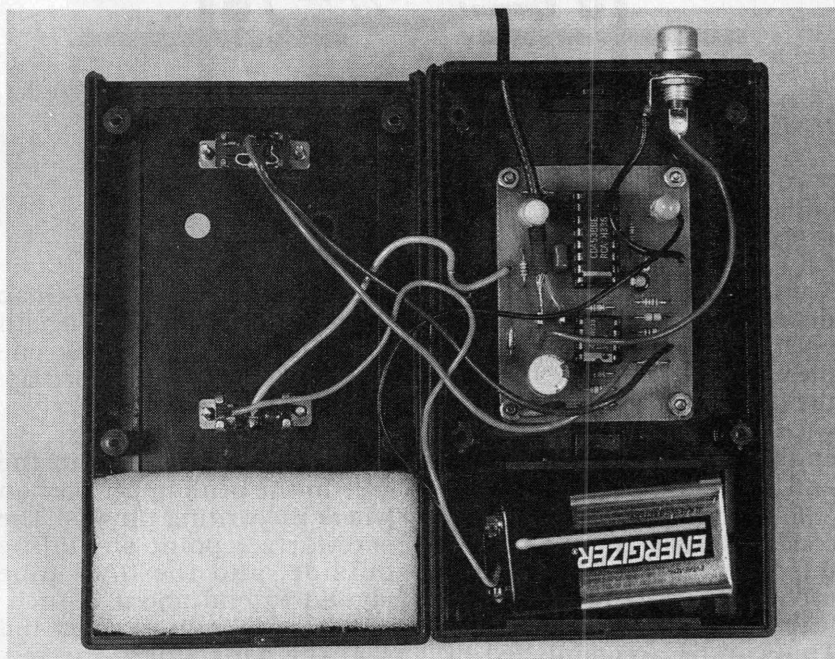


FIG. 4—THE INSIDE OF THE COMPLETED UNIT. You can use the same case (see the Parts List) or any other that will accept the board and a 9-volt battery.

mounting it if you're not using it. The rectangular holes for the switches can be made by drilling a pilot hole in the center and carefully cutting away the plastic with a sharp hobby knife. Mount the switches and solder the wires to them. Figure 4 shows the inside of the completed unit.

The IR probe

The prototype's photodiode probe case was made from a fine-point (not extra-fine point) Sanford *Sharpie* felt-tip marker pen. Figure 5 shows a cutaway view of the probe. To disassemble the pen, first pull out the writing tip with pliers. Grasp the pen's upper portion (the part that is the same color as the ink) and the pen's gray barrel. Then pull the pen apart with a twisting, bending motion. Wear rubber gloves to improve your grip on the pen. Discard the ink cartridge and wash the pen's interior with denatured alcohol to remove any remaining ink. Denatured alcohol will also remove the embossed lettering on the outside of the pen's barrel.

To make the "light pipe" that conducts light into the probe's interior, cut a small piece of $\frac{1}{16}$ -inch thick clear Plexiglas, $\frac{1}{8}$ -inch wide and $1\frac{1}{4}$ -inches long.

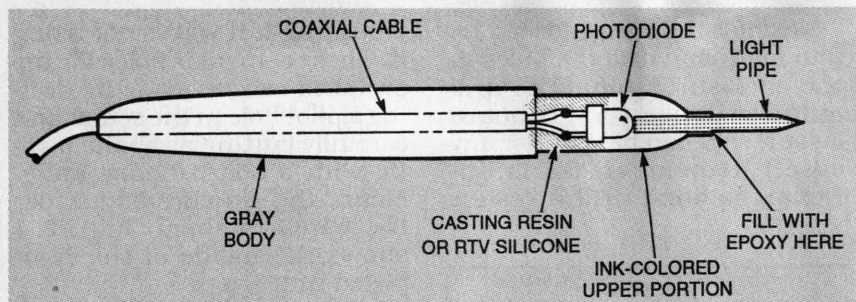


FIG. 5—CUTAWAY VIEW OF THE PROBE. The probe case was made from a fine-point felt-tip marker pen.

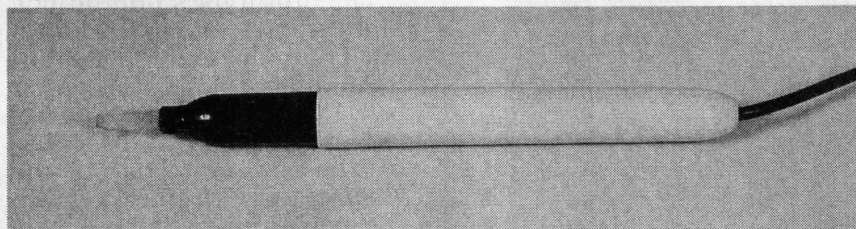
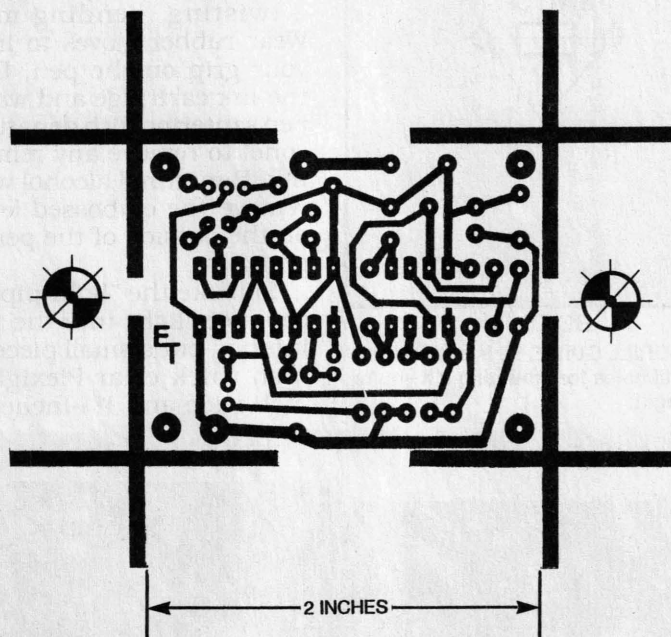


FIG. 6—THE COMPLETED PROBE. Use the pen's original cap to protect the light pipe from breakage.



IR PROBE FOIL PATTERN.

This can be accomplished by deeply scribing the sheet on both sides and clamping the piece to be cut off in a vise. Snap the piece off and cut it to length with diagonal cutters. File one end to a screwdriver-shaped tip, and then dress also clean up the

To make the "light pipe" that conducts light into the probe's interior, cut a small piece of $\frac{1}{10}$ -inch thick clear Plexiglas, $\frac{1}{8}$ -inch wide and $1\frac{1}{4}$ -inches long. This can be accomplished by deeply scribing the sheet on both sides and clamping the

piece to be cut off in a vise. Snap the piece off and cut it to length with diagonal cutters. File one end to a screwdriver-shaped tip, and then dress up the sides with the file.

Tap the light pipe into the hole in the pen's upper portion where the writing tip was. The screwdriver point should be outside, and the light pipe should extend about $\frac{1}{2}$ -inch. Mix some five-minute epoxy and seal the gaps where the rectangular light pipe enters the round hole in the pen's upper

portion.

Prepare one end of the probe's coaxial cable by stripping about an inch of the jacket off and separating the braid with an awl. Place a length of heat-shrink tubing over the cut jacket for a neat appearance. Drill a hole in the end of the marker's gray barrel large enough to admit the coaxial cable.

The photodiode's leads must be bent to extend from the center of the device. Mark the photodiode's cathode lead (it's the shorter lead), and then cut both leads down to a $\frac{1}{4}$ -inch length from the back of the device. Solder the leads of the coaxial cable to the leads of the photodiode; use the cable's braid for the cathode and the center conductor for the anode. Make sure the leads cannot short together!

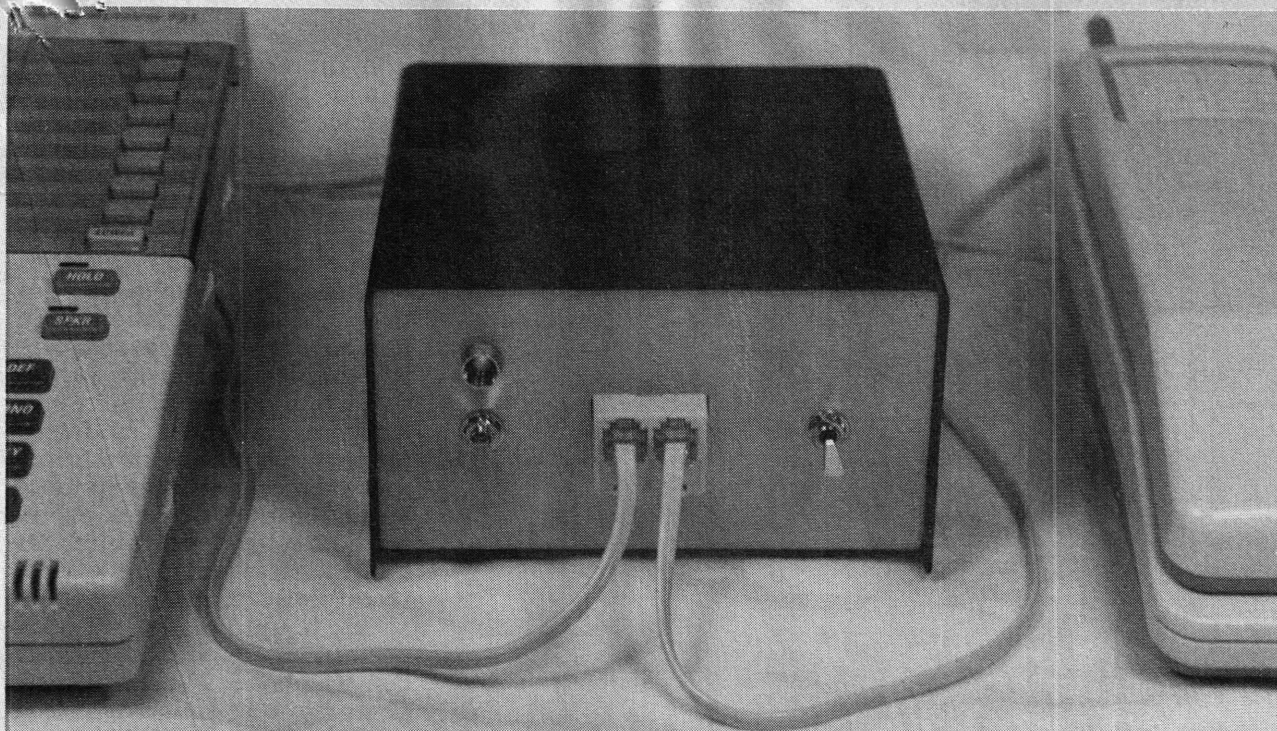
Mix about an ounce of clear casting resin (available at an art supply store) according to the directions on the package. (Alternatively, you can use clear RTV silicone sealant.) Place the probe's upper portion in a vise with the open end facing upwards. Insert the photodiode in the open end, and push it down until its lens touches the light pipe. Pour the resin (or RTV silicone) in the open end and completely fill the void, encapsulating the diode in the marker's upper body. The coaxial cable should be positioned in the center of the upper body while the resin hardens overnight.

When the resin (or RTV silicone) has cured, apply black paint to the resin around the cable. This prevents infrared light from entering through the probe's gray barrel. When the paint dries, slide the barrel over the cable, and push the two sections of the probe together. Sand the light pipe with 400-grit sandpaper to finish the surface. Use the pen's original cap to protect the light pipe from breakage when it is not in use. Figure 6 shows the completed probe.

Final assembly & testing

Pass the free end of the probe's coaxial cable through the hole in the enclosure, and

Continued on page 87



TELCO IN A BOX

***Simulate a telephone line
with the Telephone Company in a Box.***

YOU JUST BOUGHT A BRAND NEW ANSWERING machine and you want to test it out. You set it up, connect it to the telephone line, and then just sit there looking at it. You want to see and hear it work, so you drive to a pay telephone and call your home number—nothing happens. Then you rush back home, decide to read the user's manual to find the problem, and then rush back to the telephone booth to test it again.

Now suppose that instead of an answering machine, you have some old telephones lying around in your house that don't work any more. To test those telephones completely, they must be connected to your telephone line and then someone must call you. Perhaps you have a modem or a fax machine that you would like to test but don't want to pay your local copy shop \$1 a page.

JAMES E. CICON

These hypothetical examples are some of reasons why you will want to build a Telephone Company in a Box, or TCB for short, the subject of this article. It will solve all of the problems posed in these scenarios, cheaply and quickly. All of the components needed to build this project are readily available. You will not need a PC board, nor will you be required to program a microprocessor or microcontroller.

The telephone company

To understand how the TCB works, it is helpful to know what happens on your telephone line when an outgoing call is made or an incoming call is received. When a telephone handset is on-hook, the telephone line voltage is about 50 volts DC. When a handset is off-

hook, the telephone line is loaded and the voltage drops to about 7 volts DC; this voltage is detected by the telephone company's central office equipment as an off-hook condition. The central office then provides a dial tone and the equipment waits for you to start dialing. When you dial a telephone number, the central-office equipment halts the dial tone, waits to receive a valid dialed number, and then makes a connection between your telephone and the number you dialed.

When the telephone company's central office equipment has a call for you, it rings your telephone by pulsing the 50-volt DC line voltage on and off at about 20 Hz for a short period of time, then it pauses, pulses again, pauses again, and so on. This produces the ring-ring-ring effect that informs you that there is an incoming call. When

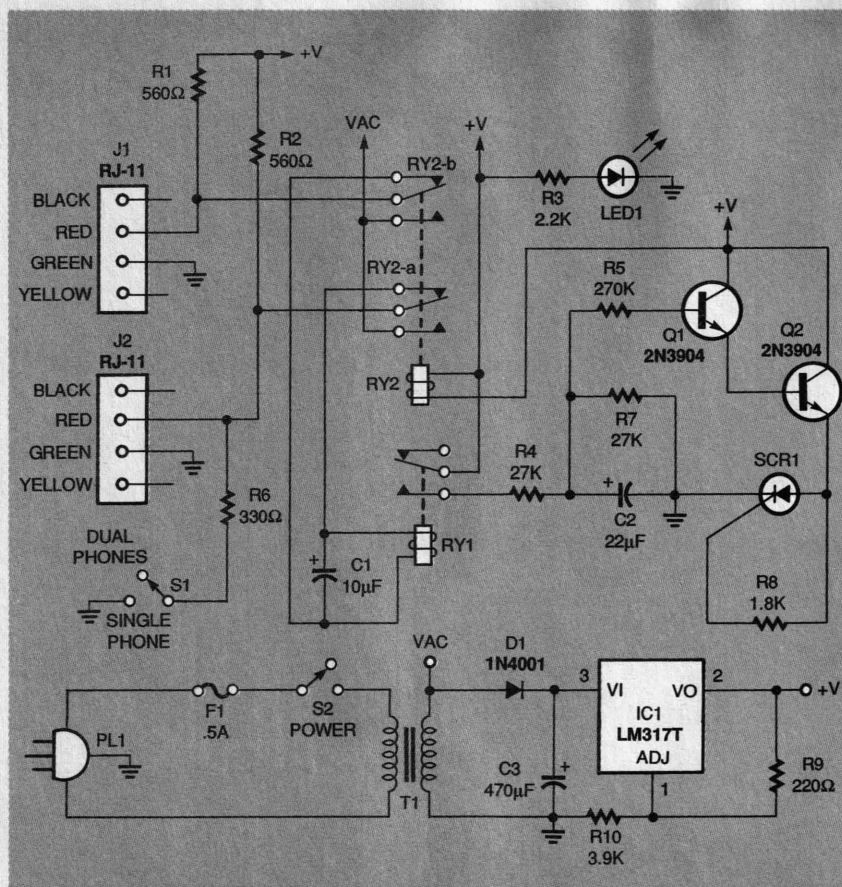


FIG. 1—TCB SCHEMATIC. When telephones plugged into J1 and J2 are both off-hook, they transmit and receive their own audio.

you pick up the handset, the telephone line voltage drops back down to about 7-volts DC. The central office equipment detects the drop in line voltage, stops ringing the telephone, and connects your telephone to the calling party.

How does TCB work?

Two telephones can be plugged into the TCB, which will then simulate all telephone-line functions. When both telephones are on-hook, the TCB supplies 24-volt DC line voltage to both telephones. Although the telephone company supplies 50 volts, the 24 volts will work well because the resistors in series with the telephones have lower values than those used by the telephone company. Moreover, 24 volts DC is both safer to work with and easier to generate.

When one telephone handset is taken off-hook, its line voltage drops to 7 volts DC. The TCB senses this and, if the

other telephone handset is on-hook, it emits a ring signal. The ring signal is a 60-Hz sinewave at about 37-volts peak-to-peak. It is applied to the line for one second and halted for one second, repeatedly, until the other handset is picked up. Although the telephone company's ring signal is a 20-Hz squarewave at 50-volts peak-to-peak, the lower ringing voltage is used for two reasons. One is for safety, and the other is because it's easier to pick the ring voltage directly from the secondary of an AC transformer than to generate it with additional circuitry.

When both handsets are off-hook, both lines are loaded to 10-volts DC. The TCB senses the condition, halts the ring signal, and connects both telephones together. Note that when the TCB is ringing the on-hook telephone, it is also sending the ring signal to the off-hook telephone. Depending on the design of the off-hook telephone, it might or might not ring. You

might also hear the ringing from the speaker of the handset as a loud buzzing. This is not a problem—just a distraction. Telephones are designed to accept a ring signal even when they are off-hook.

Circuit description

Figure 1 is the schematic for the TCB circuit. When both handsets are on-hook, resistors R1 and R2 supply power to them, and the rest of the circuit can be ignored. When both telephones are off-hook, they transmit and receive their own audio, so all that the TCB does is supply power to them through R1 and R2. If switch S1 is closed, resistor R6 simulates a telephone being plugged into J2, permitting the testing of only one telephone. The rest of the circuit can be ignored in this example.

Now consider the situation where one handset is on-hook and one is off-hook. That causes

PARTS LIST

All resistors are ½-watt, 5%.

R1, R2—560 ohms
R10—3900 ohms
R3—2200 ohms
R4, R7—27,000 ohms
R5—270,000 ohms
R6—330 ohms
R8—1800 ohms
R9—220 ohms

Capacitors

C1—10 μ F, 35 volts, electrolytic
C2—22 μ F, 35 volts, electrolytic
C3—470 μ F, 35 volts, electrolytic

Semiconductors

D1—1N4001 diode
Q1, Q2—2N3904 NPN transistor
LED1—light-emitting diode, any color
IC1—LM317 3-terminal adjustable positive voltage regulator
SCR1—200-volt, 6-ampere silicon controlled rectifier (Radio Shack No. 276-1067)

Other components

F1—½-ampere fuse
J1, J2—RJ-11 telephone jack (see text)
PL1—AC plug and linecord
RY1—SPST reed relay (12 VDC, 11 mA coil; 1-ampere, 125 VAC contacts)
RY2—DPDT miniature relay (12 VDC, 43 mA coil; 1-ampere, 125 VAC contacts)

S1, S2—SPST switch

T1—120/25.2 VAC transformer, 450 mA

Miscellaneous: perforated construction board, fuse holder, project case, wire, solder

the on-hook line to go to 24 volts DC and the off-hook line to go to 7 volts DC. The coil of relay RY1 is connected across the two telephone lines, and the voltage difference between the two lines energizes it. When the contacts of RY1 are closed, C2 charges through R4; it takes about 1 second for C2 to charge to 12 volts DC. The 12-volts DC across C2

Resistor R8 supplies gate current to the SCR, which is organized so that C2 must charge to 12 volts DC before the SCR will turn on. When SCR1 turns on, the switch is essentially closed, and relay RY2 is energized. When the voltage across C2 drops to about 2.4 volts, Q1 and Q2 force SCR1 to turn off, thus turning the electronic switch

for less than \$1. The female end was cut off, making it a line-cord. Jacks J1 and J2 were two parts of a duplex telephone jack adapter that was cut apart. Snap off the male end and solder the wires directly to the exposed leads. However, any RJ-11 jacks will work well. Mount the finished board in a metal case. Any standard project case will do. Mount jacks J1 and J2, switches S1 and S2, and power indicator LED1 on the front panel of the case. Figure 2 shows the inside of the author's prototype unit.

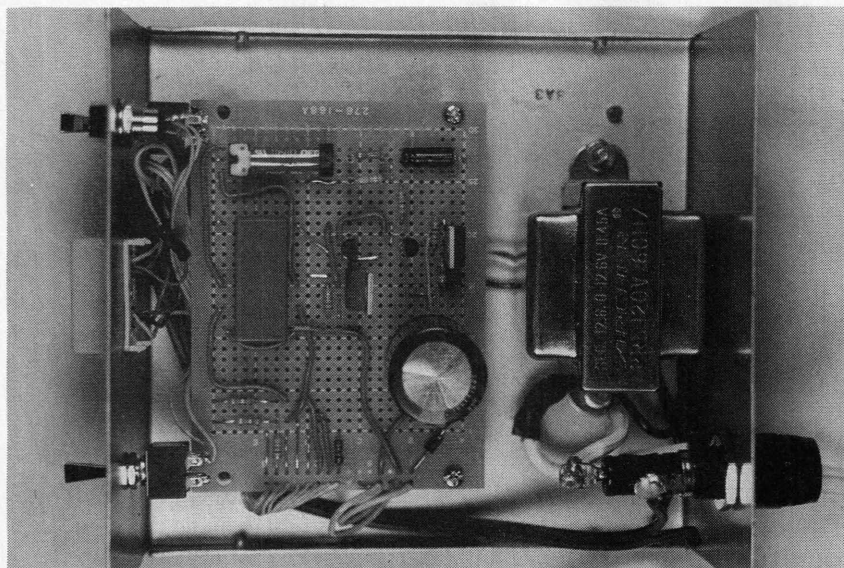


FIG. 2—THE FINISHED BOARD can be mounted in a metal case. Jacks J1 and J2, switches S1 and S2, and LED1 are mounted on the front panel.

causes a voltage-controlled switch consisting of R5, Q1, Q2, SCR1, and R8 to close, thus energizing RY2.

When RY2 is energized, RY1 is removed from the circuit and a 60-Hz, 37-volt peak-to-peak sine wave is placed on the telephone lines causing the telephones to ring. Because RY1 is removed from the circuit, capacitor C2 starts discharging through R7. It takes about 1 second for the capacitor to discharge to about 2.4 volts DC. That lower voltage level causes the voltage-controlled switch to disable RY2, removing the ring voltage from the telephone lines and putting relay RY1 back in the circuit. If one telephone is still off-hook and one is on-hook, the cycle is repeated.

The voltage-controlled switch works as follows: Transistors Q1 and Q2 are connected as a Darlington pair, which couples the voltage across C2 to SCR1.

off. Resistor R3 and LED1 provide a power-on indication for the user.

Construction

The circuit is so simple that a PC board is not necessary. Use point-to-point wiring. The power cord and AC plug PL1 were obtained from a 6 foot extension cord that usually sells

How to use TCB

The TCB is easy to use. Plug a telephone into each of the two telephone jacks and turn the unit on. If you pick up the handset of one of the telephones, the other telephone will ring. When the handset of the other telephone is picked up the ringing stops, and you can talk normally between the telephones, as if a connection had been made by your local telephone company.

If you are testing an answering machine, plug it into one jack, and plug a standard telephone into the other jack. Pick up the handset on the standard telephone, and the answering machine will start ringing and should pick up the line. You will then hear the answering machine's outgoing message in the handset you are holding. The same procedure applies to the testing of modems, fax machines, or virtually any other telephone device whose operation you want to verify. Ω

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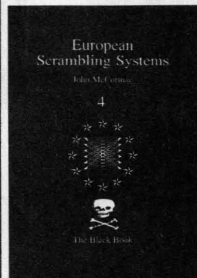
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"All channels" FM transmitter

I've had several helpline requests for this one, as a paging override for a factory or plant cable system, or as a "please buy this house" realtor's message one that is received on an FM tuner set

to any station. I can also see several emergency vehicle traffic-clearing applications for this as well.

I don't have a plug-and-go answer, but I do have a fairly simple paper design that *might* work, one that uses more of that *Fourier series* I have been looking at. Any decent FM receiver will exhibit the *capture* effect: When two competing signals are being received, and if one of them is even slightly larger, then *only* the stronger signal shows up as output audio, interference free. The capture effect happens because of the receiver's limiter circuitry. So, that's a good starting point. If your nearby signal is stronger than a distant one,

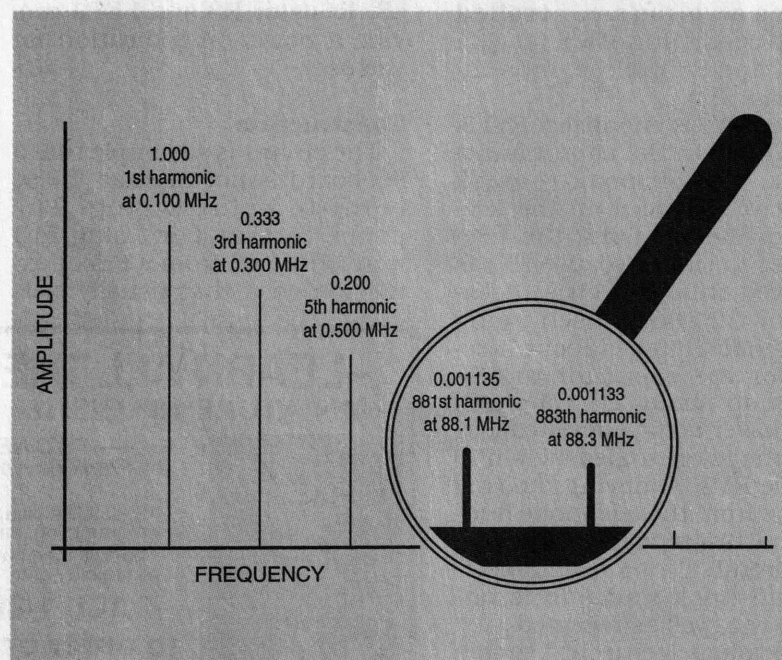


FIG. 1—A FAST RISE 100 KHZ SQUARE WAVE has harmonics well into the FM band. The harmonics match existing FM station channels.

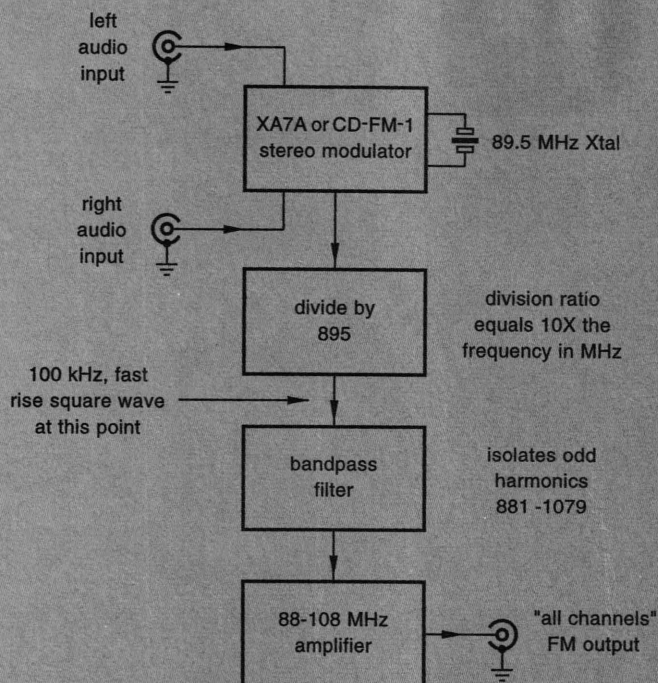


FIG. 2 — ONE POSSIBLE SCHEME for an "all channels" FM generator.

then only the stronger nearby signal will be received.

I've written about stereo broadcasters a few times in past columns. The best starting point is the *Robm* BA1404 stereo chip. The BA1404 makes a dandy low-cost stereo multiplexer, but is a terrible transmitter. Frequency drift and tuning accuracy slashes range or outright stops reception on digitally synthesized receivers, especially with car radios.

The one- or two-channel solution is to use the BA1404 to drive a crystal-stabilized FM transmitter. But linear frequency modulation of a crystal oscillator is more than a little bit tricky. A lab full of obscure test equipment, a lot of math, and years of experience are needed before you can even think about doing it.

I've explained how the *Pioneer* CD-FM-1 or the *Sony* XA7A auto CD adapters give you an off-the-shelf design solution. But the real problem here is not to transmit on a single channel, but to create *identical* signals on *all* the FM channels at once. Those FM channels run from 88.1 through 107.9 MHz, with an 0.2-MHz channel spacing.

Figures 1 and 2 show how the Fourier Series "comb transmitter" approach might handle this for us. A square wave has a fundamental frequency plus 1/3 the third harmonic, 1/5 of the fifth harmonic, and so on. So, start with a single-channel transmitter using one of those CD car adapters. Divide that down to precisely 100 kHz, making sure the final stage is an ultra-high-speed divider with a rise time better than two nanoseconds.

At this point, you have an FM square wave with lots of harmonics. For instance, there will still be two millivolts or so of the 881st harmonic at, of all places, 88.1 MHz. And, surprise, surprise, there is also an 883rd harmonic at 88.3 MHz, an 885th harmonic at 88.5 MHz, all the way up to the 1079th harmonic at, wonder of wonders, 107.9 MHz. Sure, the high-end "stations" will be a tad weaker than the low-end ones, by about 20 percent or two decibels or, from the center of the band, plus or minus one percent, or one decibel either way. So, just add some gain and a bandpass filter to build up an all-channel FM broadcaster. Those one-dollar MAR

chips from *Mini-Circuits Labs* should work just fine for predrivers in the circuit.

To be sure, you'd have problems getting this one to behave in the real world. A wide dynamic range would be needed, as well as thorough shielding. To prevent instabilities, you might want to mill each stage out of a block of solid aluminum. Or, at the very least, use stripline and double ground planes all the way around. There's also some very restrictive FCC regulations on this sort of thing. Transmitting unwanted messages to anyone, ever, is a definite no-no. But neither of these should be a problem on any private cable system. One solution to a "buy my house" application is to use several ultra-low-power curbside sites. Or perhaps go to some sort of a "lossy line" leaky cable that acts as a distributed antenna.

Filter fundamentals

An electronic filter can be defined as a frequency-selective network. A filter can emphasize wanted information while rejecting noise. *Low-pass*

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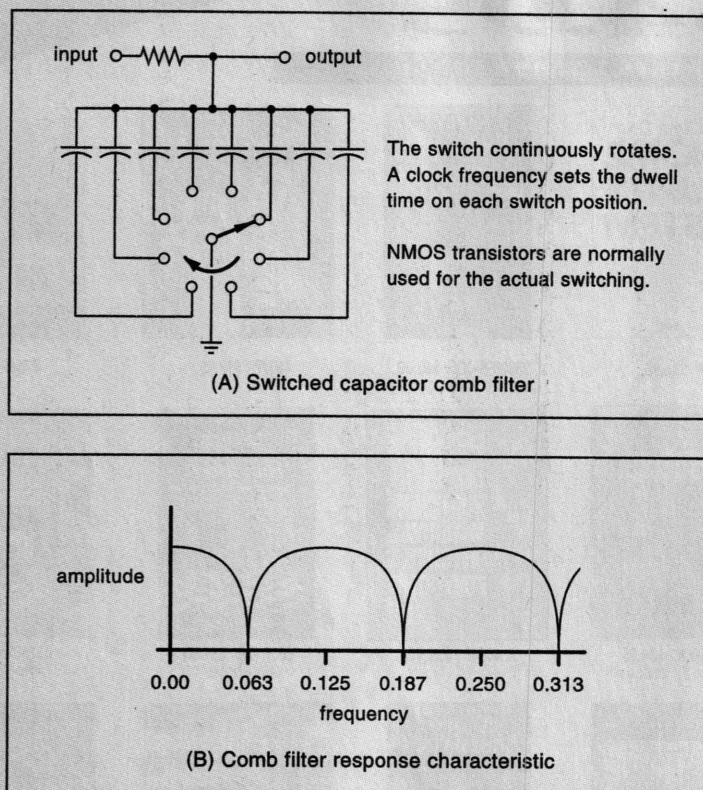


FIG. 3—COMB FILTER BUILT UP by using switched capacitor techniques. Lowpass, bandpass, and highpass versions are also possible.

filters sustain the lower frequencies but obstruct higher ones. One example of a low-pass filter is the treble control on any hi-fi. Noise-reducing capacitors connected to ground in a power supply line are also low-pass filters, often called *bypasses*.

Bandpass filters select midrange frequencies while rejecting the higher and lower ones—the tuning dial on a radio, for instance. Audio equalizers are made from overlapping bandpass filter circuits, each of which can be adjusted independently.

High-pass filters block lower frequencies but emphasize the higher ones. The bass control on a hi-fi is one example of a high-pass filter. By adjusting its cutoff frequency, you can selectively de-emphasize any of the lower musical notes.

Actually, there is no such thing as an electronic high-pass filter. One of these would also be able to transmit microwaves, heat, light, and cosmic rays as well. The frequency response of any electronic circuit usually sets an upper limit, making all high-pass

filters into bandpass ones.

Similarly, a true low-pass filter must respond down to DC, including any circuit bias or offsets. Audio low-pass filters add a large decoupling capacitor to block DC, and thus become a bandpass filter in the process. Other more specialized filter types include *band-stop* filters that reject a range of frequencies, *notch* filters that are just very narrow band-stop filters, and *delay networks* that adjust phasing at chosen frequencies.

There are several ways to build a filter. *Classic* filters use fixed inductors and capacitors. *Transmission-line* filters are just the UHF or microwave equivalent of classic *L-C* filters. The big difference is that the inductance and capacitance is *distributed*, rather than *lumped*. *Active* filters use combinations of resistors, capacitors, and op-amps instead. Advantages include low cost, high gain, and easy tuning.

Mechanical filters make use of the physical resonance of a crystal, ceramic resonator, or magnetostrictive rods. *Surface-wave* filters use con-

structive and destructive interference of sonic waves on a piezoelectric surface. A *switched-capacitor* filter is a hybrid analog and digital scheme—more on this shortly. Finally, *digital* filters will treat the filter problem as a group of numbers with digital signal processing, FFT transforms, or even wavelets.

The *order* of any filter determines how strong or how powerful it is. For instance, a *first order* low-pass filter should, at best, fall off at six decibels per octave, cutting its response in half as the frequency doubles.

A *second-order* low-pass filter can have a response that drops 12 decibels per octave. An *eighth-order* filter drops down at a rate of 48 decibels per octave. This gets real tricky on band-pass designs. For instance, a second-order bandpass filter might drop off rather steeply just outside its pass-band, but its ultimate high- or low-frequency falloff will be six dB per octave. The initial falloff rate is set by the *Q* of the filter which, for a second-order section, ends up as the inverse of the bandwidth.

The *style* of a filter is set by the underlying math used to determine its response. Filters using *Bessel* math provide the flattest time delay but a poor amplitude response. *Butterworth* filters provide the flattest passband amplitude. *Chebyshev* filters provide faster amplitude drop-offs, but at the price of lumpy passband ripple. *Elliptic* filters will add one or more notches just into the stopband, and give the steepest possible falloff, but these suffer from far poorer ultimate signal rejection.

Actually, these strange names are part of a *continuum* of possible filter responses. In an active filter, all you have is a stack of cascaded first- and second-order sections. Each one has a *cutoff frequency* and a *damping factor*. By suitably changing these values, you can create any filter from Bessel on up through elliptic.

In general, you can't do nearly as much with a filter as you'd like to. Increasing the order adds cost, noise, linearity, dynamic range, shielding, losses, tuning, stability, and response problems, besides totally trashing the transients and time delays.

One nasty filter response is called *group delay*. For instance, if a modem tone frequency for a one has a different group delay than that for the zero, then at times you'll get a one, a zero, neither, or both.

In general, digital filters let you do much more, much better than analog filters. For instance, a filter with a "brick wall" or "reject everything" response is impossible with analog designs, but can be created easily with digital techniques. Group delay is also more easily dealt with, and so are the elegant new *constant phase* designs.

Sadly, digital filters are currently restricted to operating on big amplitudes and low frequencies, or to non real-time uses. And they have their own set of problems, not the least of which is *aliasing*, or generating nasty in-band artifacts from out-of-band signals.

By using wavelets or fast Fourier transforms, you can find the full spectrum of a waveform rapidly. Analog filters must instead be swept slowly. You can soon expect to see a revolution in radio receivers. Start at the front end, downconvert, and then digitize. From there, do *everything* digitally. Several technical notes on this are offered by *Harris Semiconductor* and *Analog Devices*. You'll find much more on filtering in my *Active Filter Cookbook*, now in its brand new second edition and *seventeenth* printing. See my nearby *Synergetics* ad for information on its availability.

Switched-capacitor filters

There's an interesting hybrid filter scheme based on both analog and digital technology. This is known as a switched-capacitor filter. The basic idea is shown in Figs. 3 and 4. The switch goes 'round and 'round at a clock rate that is a multiple of the desired center frequency. Now, for a given switch position, there's only one capacitor in the circuit. This capacitor should either increase or decrease its charge according to the *difference* between itself and the input signal. Each capacitor thus retains a *sample* of the recent previous input history.

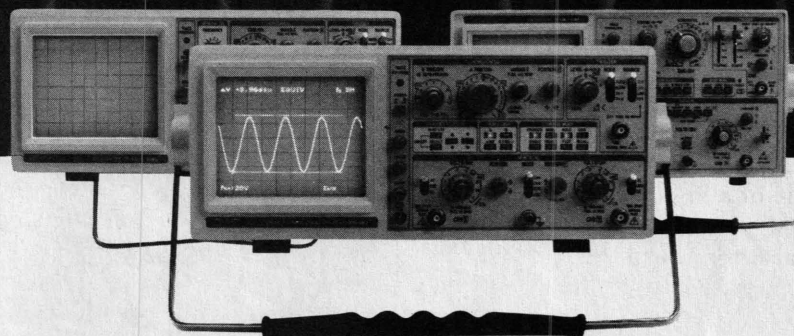
Now for the neat part: At DC or

very low frequencies, the capacitors will pretty much track the input. As the frequency increases, the capacitors may charge or discharge. At a frequency of exactly *one-half* the one-trip-around rate, any capacitor will charge on one pass and discharge on the next, giving a zero output. At a frequency of precisely the one-trip-around rate, any capacitor charges identically on each pass and gives you a *maximum* output.

This simple example forms a *comb* filter. The first null frequency will be

determined by the switching rate and the number of capacitors. In this example, the first null frequency is at 1/16th the data rate. Comb filters can reject a fundamental and *all* of its harmonics. One major application for comb filters is separating NTSC color and brightness signals. Different combinations of resistors, switches, and capacitors could build high-quality low-pass, high-pass, band-pass, or band-stop responses. Switching is normally handled by CMOS analog switches.

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Advantages of switched capacitor filters include precise tuning over a wide frequency range, plus lower cost and small size. And because everything is done on the silicon chip, there is no need for any external capacitors.

Dozens of switched capacitor filter chips are now available from several sources, some approaching a dollar each. Our resource sidebar for this month shows you several of the main suppliers involved. Of these, check into *Maxim* and *Linear Technology* first. Free samples of these filter chips

are available from Maxim.

Sadly, there have been quite a few false starts involving these switched capacitor filters. Early versions had excessive noise floors, caused by the transient feedthrough of the clock. That happens because of the gate-to-source capacitance in any CMOS device. The dynamic range is set at the low end by the noise floor, and at the high end by clipping, distortion, or other nonlinearities. Today, these glitch and noise-floor problems have been minimized. Still, though, switched capacitor filters are strictly

limited to larger signals. They are not at all suited for microvolt-size inputs. Typical devices are limited to ultrasonic or lower frequencies. But they sure are a useful tool.

A narrow bandpass filter

There is this interesting new LTC 1164-8 chip from *Linear Technology*. This device creates an easily tunable and ultra-narrow eighth-order elliptic bandpass filter. The bandwidth is one percent of the center frequency. Cost is around \$24 in single quantities. Figure 4 shows the details. The chip goes ahead of an op-amp in such a way that the generated noise floor is independent of the gain. With a single 5-volt supply, the current draw is around two milliamperes and the maximum center frequency is 4 kHz. The input clocking frequency is 100 times the center frequency. Output noise is half a millivolt. The stopband attenuation is fifty decibels or so. Careful ground-plane shielding is required. Be certain to carefully read the LTC1164 data sheet.

Gain is adjusted by changing the input resistor. Note that the *smaller* the resistor, the *higher* the gain. A gain of 1000 is possible, but you will get the best dynamics at unity gain. To use the circuit, set the gain by picking the input resistor, apply an input signal you want filtered, and a 100~ square-wave clock derived from any old TTL or CMOS source.

For quick and dirty experiments, simply use one sixth of a 74HC14, a variable resistor, and a capacitor to make a square-wave clock source. One obvious use for the chip is for checking real-world harmonic performance of the magic sinewaves I looked at a few columns ago. But one major warning over *any* narrow band filter: *If your signal is changing rapidly, most of it will be missed!* That happens because the high Q of the filter takes a long time to build up a useful result. It's sort of the same as whapping a large pendulum with a hammer. Nothing much happens with one ping. Its only when hundreds of well timed pings occur that oscillation starts. Thus, narrow-band filters must be very accurate and tuned very *slowly*. And they will work only with very

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slowly changing signals.

The data sheet doesn't mention it, but I suspect that the LTC1164 also responds to the second and higher harmonics of your center frequency as well. This is called *aliasing*. Therefore some external prefiltering appears to be a good idea. This device seems ideal for a lot of telecommunication and midrange audio applications, especially for vibration studies and

harmonic analysis. But, for ultra-low frequencies (such as seismography or brainwave research), digital FFT or wavelet techniques are the only way to go these days.

Two contests

Let's have a pair of contests this month. Either (A) Add to our dialog

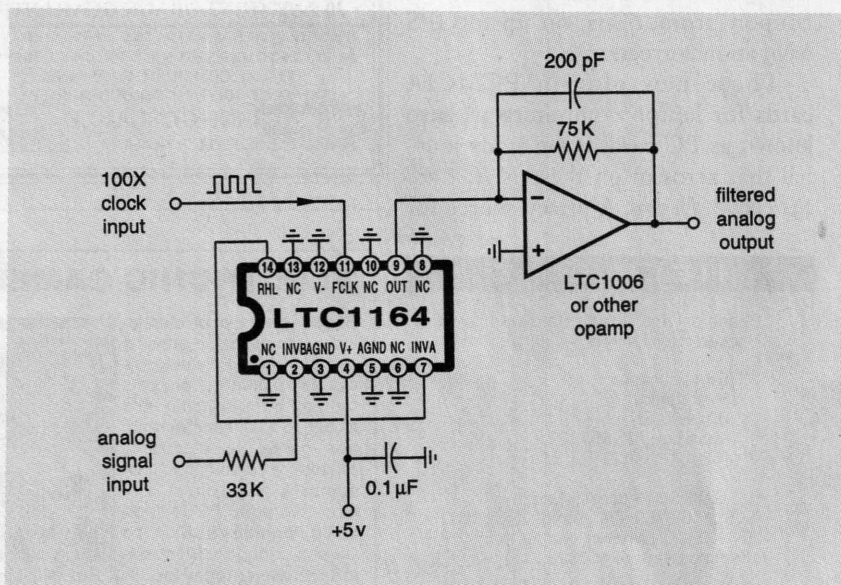
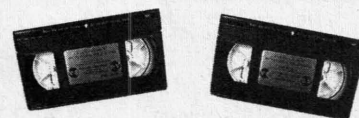


FIG. 4—AN ULTRA SELECTIVE switched capacitor bandpass filter. The center frequency can be tuned by adjusting the digital 100× clock.

TV/VCR TECHNICIANS



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VCR/TV Repair TECHDISK

This software database contains over 4800 VCR & 2100 TV problems, along with practical repair suggestions for each, compiled from case histories. User friendly and menu driven. Add/delete your own problems too. Annual updates available for a nominal charge. (IBM) Order# 56-130. \$69.95 2 disk set. Also available separately for \$39.95 ea.

VCR Model SuperCross

This software database allows you to cross between various makes and models. For example, enter a VCR model# and find out what other VCRs are electrically and mechanically the same. This info is helpful for parts/service manual ordering. Menu driven and user friendly. (IBM) Order# 56-300. \$39.95

Directory/Electronic Manufacturers

This priceless software contains hundreds of electronic manufacturers telephone and fax numbers, technical assistance and toll free numbers, and BBS numbers. Addresses are also included. Includes manufacturers of VCR, TV, Microwave, Computer, Audio, and more! (IBM) Order# 56-107. \$29.95

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books on the PCMCIA standard is *Sycard*. One source of cards is *Quatech*.

Bargain hydraulics are offered by *Roberts*. Unusual modelmaking parts are available at *Special Shapes*.

The magic sinewave stuff that I looked at a few columns or ago has now been improved and upgraded as *MAGICSIN.PDF*. I've got actual chips and working hardware. Co-developer packages are newly available for this emerging opportunity.

For most small-scale startups most of the time, patents are certain to end up as a net loss of time, energy, money, and sanity. Find out why this is true in my *Case Against Patents* resource package. An autographed copy of my *Incredible Secret Money Machine II* is included in the patents resource pack.

My usual reminder that most of the resources mentioned appear in the *Names & Numbers* or that *Switched Capacitor Filter Resources* sidebar. I've managed to wrangle a special ten free hour *GENie* PSRT trial per the *Need Help?* box. **EN**

on "all channels" FM schemes, or (B) Show me an unusual new use for an ultra narrow, switched capacitor bandpass filter.

There'll be a dozen or so of my *Incredible Secret Money Machine II* books going to the better entries. Plus an all-expense-paid (FOB Thatcher, AZ) *tinaja quest* for two going to the very best of all. Please send your *written* entries to me here at *Synergetics*, rather than to *Electronics Now* editorial.

New tech lit

An update on those active noise-cancellation topics we looked into several columns back: I've uploaded a tutorial and resource directory as *NOISCNCL.TXT*. There's also a new *Active Sound and Vibration Control News* with free sample copies available.

More on FM radio RBDS services: There is now an Internet forum up at <http://pcbf1131e.util.ch/~uer/rdsh000>

.htm. As always, *Radio World* is the best place to get RBDS information. Hundreds of stations now provide RBDS services, ranging from song title and artist, traffic and weather, coupon promotions, on up to GPS navigation corrections.

Those new plug-in PCMCIA cards for laptop computers are also known as PC cards. One trade journal that zeros in on these is *IC Card Systems & Design*. A pricey source for

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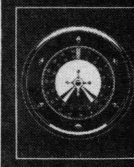
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Electronic Games



Using Lasers to Make Holograms



ONE OF THE MOST FASCINATING ASPECTS ASSOCIATED WITH LASER RESEARCH IS ITS ABILITY TO PRODUCE PHOTOGRAPHS WITH THREE DIMENSIONS. ACTUALLY, IT IS THE LASER'S PROPERTIES OF COHERENCE AND PHASE THAT MAKE THESE

photographs appear to be 3-D. In a normal photo, the direction and intensity of light on any given part of the subject is recorded on the film. With the laser photo, or hologram, the distance of each part of the subject from the film is also recorded. Here's how:

Since different areas of the subject are either closer or further away from the film, the light reflected from the object strikes the film at different phases. That reflected light is known as the object beam. A second light source, or reference beam, is projected directly onto the film. Since no reflections take place, the phase of that light is uniform. By combining the two beams on the photosensitive plate or film, the mismatches in phase set up an interference pattern, which creates the three-dimensional effect.

One restriction in viewing holograms is that the viewing light must be of the same frequency as the exposure light. By using the same laser for both, that problem is eliminated. The image produced will appear in the same color as the light source. Since we are using a He-Ne laser, the image will be in red.

With theory out of the way, let's turn our attention to the equipment we will need. That includes the laser, some optics and mirrors, film and developer, and a special layout table. Once you have those, you have everything that you need to shoot, process, and display holograms. The laser and optics should be old friends by now. We'll take a careful look at the

film and processing chemicals later. That leaves the assembly table.

The Assembly Table

The major enemy of good, clean, holograms is vibration. Long exposures are needed. Therefore, it is essential that the components of a holographic arrangement be kept as motionless as possible while the picture is being shot. Vibration can be caused by everything from accidentally bumping the table, to passing trucks, to even thunder. That's why the assembly table must be sturdily built and designed to dampen all unwanted movement.

The first consideration is where to place the table. Pick an area on a ground floor or basement that is part of the foundation. A garage with a solid concrete floor is a good choice. Also, keep in mind that the exposures have to be made in total darkness (except for the light from the laser); the room must be selected or modified to accommodate that, or you will need to build a light-tight hood for the table.

Once you have selected a suitable location, erect the table supports, a maximum of 24 inches high, in the appropriate spots. Obviously, that is dictated by the size of the table. Cinder-block or concrete-block legs with squares of heavy carpet on top work quite well. Situate them at each of the corners, and inset them about 3 inches from each side. If you are building a large table, place a

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This article deals with and involves subject matter and the use of materials and substances that may be hazardous to health and life. Do not attempt to implement or use the information contained herein unless you are experienced and skilled with respect to such subject matter, materials and substances. Neither the publisher nor the author make any representations as for the completeness or the accuracy of the information contained herein and disclaim any liability for damages or injuries, whether caused by or arising from the lack of completeness, inaccuracies of the information, misinterpretations of the directions, misapplication of the information or otherwise.

fifth leg to support the center. If additional legs are needed for total stability, add them.

While the table can be a almost any size, a couple of considerations need to be kept in mind. It does need to be large enough to accommodate all the components of the assembly, and it needs to be as heavy as is practical to help smooth out vibration. Figure 1 shows a 50- by 20- by 10-inch table that will weigh in excess of 125 pounds when completed. It provides enough room for most experiments, as well as all the stability you are likely to need.

The diagram is self-explanatory, but here is a step-by-step discussion of the construction. First, the primary case, measuring 50 by 20 by 10 inches (outside dimensions), is built from ½-inch plywood. It is probably wise to place this box on the support legs at this time, as the total weight will start adding up very quickly. Next, inflate to firm, but not full, three 7- by 16-inch diameter inner tubes. Place them, side by side, in the bottom of the main case. The rubber tubes act as the primary damper for vibration. Now

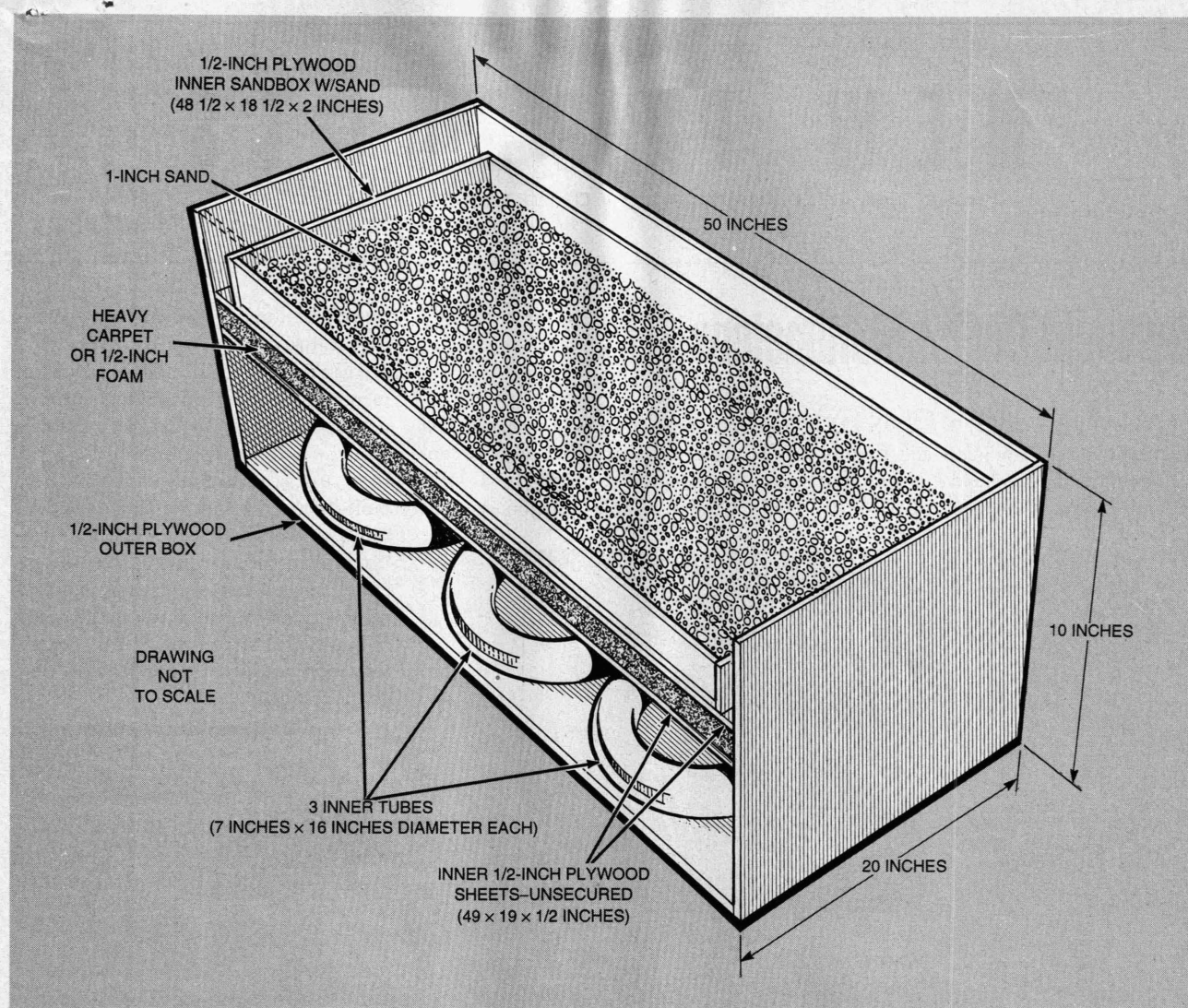


FIG. 1—A HOLOGRAPHY ASSEMBLY TABLE must be solidly built and positioned in a vibration-free area. Long exposure times are required, so no movement can be tolerated.

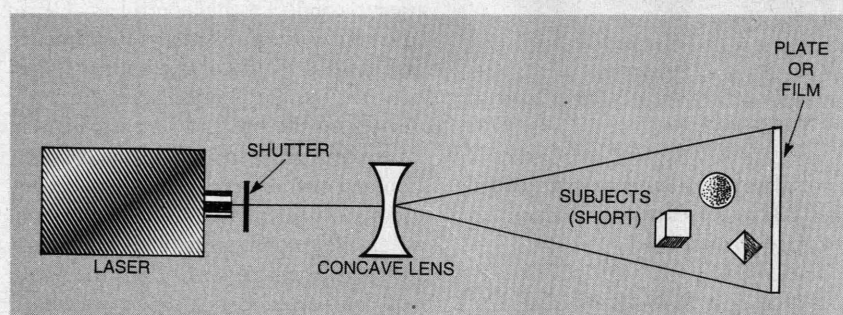


FIG. 2—USE THIS SINGLE-BEAM HOLOGRAM layout to make transmission-type holograms when shorter subjects are involved.

lay a 49- by 19-inch sheet of 1/2-inch plywood on top of the inner tubes. Heavy carpet or 1/2-inch foam rubber, also measuring 49 by 19 inches, is put in place and another 1/2-inch sheet of plywood goes on top of that. Both plywood sheets are cut to fit tightly, but must not bind with the inside of the main box. A second 1/2-inch plywood box, measuring 48 1/2 by 18 1/2 by

2 inches, is built and placed on top of the second "free-moving" sheet. Fill that box with sand to a depth of about 1 inch. The sand serves two purposes. It contributes about 80 pounds of weight to the table, and it provides a flexible and stable way to position the assembly components. With everything in place, smooth the sand to provide an even surface and your table is

ready for action. Now, let's look at how to make the holograms themselves.

Two Types Of Holograms

Holograms come in two basic varieties, transmission and reflection. With transmission holography, the subject is placed in front of the film and a single, diffused beam acts as both the reference and object beam. With short subjects, the film can be directly behind. With taller objects, it needs to be positioned to one side. The setup for shorter objects is shown in Fig. 2; the one for taller subjects is shown in Fig. 3. Both are for single-beam transmission holograms.

The reflection method works just as the term implies, with the film or plate in front of the subject and facing away from the diffused beam, instead of into it. Here, again, a single beam acts as both object and reference. The differ-

continued on page 74 **21**

EASY POCSAG SIGNAL DECODER

*Learn how
alpha-numeric pagers
receive messages with this simple interface
and computer software.*

ROBERT B. WHITAKER KI5PG

Have you ever thought about how those alpha-numeric pagers work? Have you heard erratic buzzing and beeping digital signals while scanning across the VHF or UHF bands? Message services for small, portable pagers have become as widespread as cellular telephones.

But is it possible to decode the pager messages flying around the airwaves? It is easy to do from a technical point of view, but the information contained in the radio signals is a completely different matter from a legal point of view. The willful intercepting a non-broadcast-type signal meant for private communication other than a tone-only signal is a violation of law and carries the same penalty and criminal status as intercepting cellular-telephone calls.

That said, a scanner, a simple interface circuit, a personal computer, and a shareware program available through the Internet are all you need to set up your own pager-signal monitor. The monitor described here will decode 512-

1200-, and 2400-baud data streams. An additional feature of the software is that it can be configured to only decode signals sent to your own pager or a pager for which you have permission from the owner to receive. If the monitor is used solely on ham bands for monitoring ham pagers, any legal limitation might not apply. To be safe, always check with a legal advisor before using the decoder in your area.

What is POCSAG? POCSAG stands for Post Office Code Standardization Advisory Group. That advisory group has established the standard signal code and transmission protocols now in use by the vast majority of pager services. POCSAG is sent by frequency-shift keying (FSK) an FM-carrier wave at ± 4.5 kHz. A list of commonly-used frequencies for POCSAG signals are shown in Table 1. Check with your pager service to find the actual frequency they use.

The decoder will work with regular speaker audio from an external speaker jack or headphone jack, but will work much more reliably

WARNING!

Please note that unauthorized electronic communications interception is illegal under Federal and State Law. In addition Federal law renders illegal the intentional manufacturing, assembling, possessing or selling of any electronic, mechanical or other device, knowing or having reason to know that the design of such device renders it primarily useful for the purpose of surreptitious interception of oral or electronic communications. Federal law imposes both civil and criminal penalties for violations of the applicable statutes. Thus, the use of the POCSAG Signal Decoder described in this article is intended for and should be restricted to educational, scientific and/or informational purposes. This is not intended to constitute legal advice and readers are advised to obtain independent advice as to the propriety of their use thereof based upon their individual circumstances and jurisdictions.

with raw de-emphasized audio taken directly from the discriminator of the receiver or scanner. Scanners or radios that are 9600-packet ready should be useable without any further modifications. Other scanners or radios that don't have an audio or discriminator out-

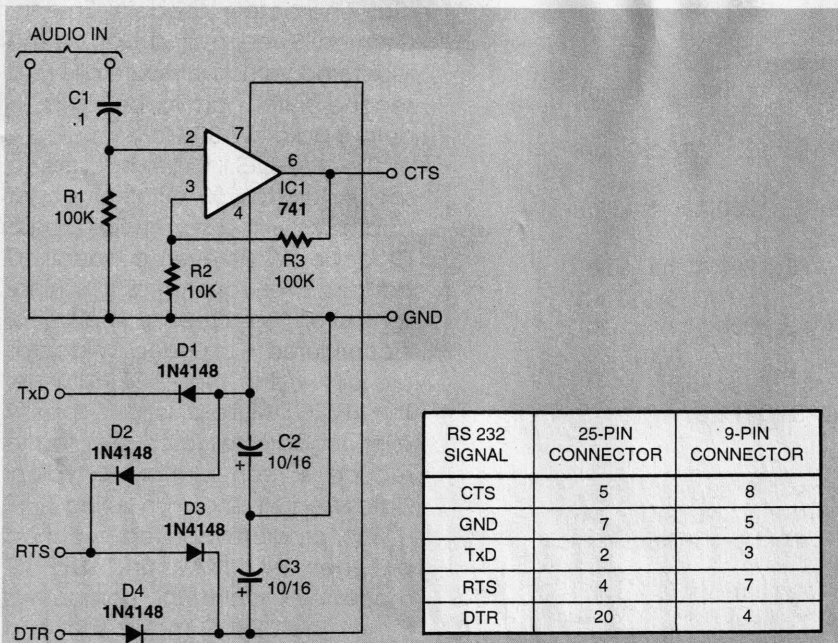


Fig. 1. The interface circuit for the POCSAG decoder is simply a comparator that takes the demodulated audio from a scanner or radio receiver and changes it to digital pulses that can be read by the software. Power for the circuit is derived from the unused pins on the serial port.

put might need to be modified in order to pick up the signal directly from the discriminator output. If you do not know where the discriminator output is on your radio, and a schematic diagram of your radio is not available, a good place to test as a pick-up point is the high-side lead on the squelch-control knob.

The Hardware. The primary component in the interface circuit of Fig. 1 is a 741 op-amp available at Radio Shack and just about all other major electronics parts vendors. Power for the op-amp can be supplied separately, or could be easily supplied from the unused Tx, Rx, and DTR pins from the computer's RS-232 serial port. Diodes D1 to D4 act as a type of bridge rectifier, making sure the supply pins of IC1 are only connected to the proper voltages from the RS-232 pins.

In this circuit the 741 op-amp is used as a comparator, converting the signal from the receiver into the ± 10 -volt signal necessary to drive the RS-232 CTS (or DSR) input. That comparator has a positive feedback network, giving hysteresis that helps to recover the data from a receiver's audio output. The level of

hysteresis, set by the R2 and R3, can be adjusted for best reception. A 100,000-ohm potentiometer could be substituted for the hysteresis network, allowing the circuit to be fine tuned. If a direct discriminator output is available from the receiver no hysteresis might be necessary; in that case, R2 could be omitted.

Construction. The circuit is simple enough to be built on a perfboard in a few minutes. An alternative is to purchase either a complete kit or an assembled and tested interface from LQpht Heavy Industries (see Parts List), a small group of Boston-area experimenters with a deep interest in electronic hardware. Note that the group has a "zero" in their name and not a capital letter "O".

If you opt for the kit, it includes a high-quality, silk-screened, double-sided PC board with plated-through holes measuring about one-inch square. Although the construction is not difficult, that circuit board (not shown here), which is designed to fit inside the hood of a DB-25 pin connector, is rather small and there is little margin for excess solder. The kit includes all documentation for the project, as well as a shareware ver-

PARTS LIST FOR THE POCSAG SIGNAL DECODER

IC1—LM741 op-amp, integrated circuit

D1—D4—1N4148 silicon diode

R1, R3—100,000-ohm, 1/4-watt, 5% carbon resistor

R2—10,000-ohm, 1/4-watt, 5% carbon resistor

C1—0.1- μ F, ceramic-disc capacitor

C2, C3—10- μ F, 16-WVDC, electrolytic capacitor

Socket for IC1, PC board, DB-25 or DE-9 female connector, hardware, wire, solder, etc.

Note: The following items are available from: LQpht Heavy Industries, POCSAG Project Division, P.O. Box 990857, Boston, MA 02199-0857; Complete kit with registered software, \$59.95; Assembled and tested unit with registered software, \$89.95; Complete kit with unregistered software, \$19.95. Please add \$5.05 for shipping and handling for each unit ordered. All payments are to be in US funds. Accepted forms of payment are cash, check, or money order only. Please make all checks payable to LQpht Heavy Industries. No COD orders will be accepted. Shipment is via first-class US mail to anywhere in the world. Massachusetts residents please add 5% sales tax. Allow four to six weeks for delivery.

sion of the software to decode the POCSAG signals on a single 3-1/2 inch PC disk. The interface kit includes some assembly instructions in the README.1ST file, although the kit designers assume that the builder has some experience in soldering and electronics experience. The same information is available through the Internet by visiting the POCSAG Web page at <http://www.lqph.com/~kingpin/pocsag.html>.

Incidentally, that Web page is quite interesting. It features a high quality image of the front and back of the interface board. If you look closely, you will notice that the background wallpaper is actually a schematic diagram of the decoder interface shown here.

Whether the interface is bread-boarded or built from the available interface kit, a fine tip, well tinned, soldering iron is essential. A fine touch and soldering experience are also beneficial. Poor soldering will undoubtedly contribute to poor results.

LISTING 1

```
20:13:03 04/18/96 ***** LOGGING STARTED *****
20:13:17 RIC: 0546426 FUNC: 0 RATE: 1200 Alpha (auto):
;;TEST POCSAG PROTOCOL
20:13:18 RIC: 0546426 FUNC: 0 RATE: 1200 Alpha (auto):
THIS IS AN EXAMPLE=OF LOPHT
20:13:18 RIC: 0546426 FUNC: 0 RATE: 1200 Alpha (auto):
HEAVY INDUSTRIES POCSAG
20:13:18 RIC: 0546426 FUNC: 3 RATE: 1200 Alpha (auto):
PAGING DECODER.==;
20:13:20 RIC: 0546426 FUNC: 0 RATE: 1200 Alpha (auto):
u0+D90f
20:13:20 RIC: 0546426 FUNC: 2 RATE: 2400 Tone
20:13:46 04/18/96 ***** LOGGING STOPPED *****
```

```
C2519A9B 7A89C197 7A89C197 7A89C197 7A89C197 7A89C197 6A7BF468 A25417F9
945C752B 7A89C197 7A89C197 7A89C197 7A89C197 7A89C197 66B4485B 8A89B23D

C35949FD C4199826 7A89C197 7A89C197 7A89C197 7A89C197 5EB561D8 D56594E2
83319AB4 7A89C197 7A89C197 7A89C197 7A89C197 7A89C197 7A89C197 7A89C197

7558A2DC A78C9623 782C5626 A41224BC F67708D7 39BED7F6 E199996F 7A89C197
78B31399 D715C277 A7719E3A 7A89C197 741A733F 98E24D4B 8E199B14 7A89C197

7A89C197 7A89C197 7A89C197 7A89C197 7A89C197 7A89C197 7A89C197 7A89C197
7A89C197 7A89C197 7A89C197 7A89C197 51761386 A50198D2 7A89C197 7A89C197

4A8B5FFD A388D69A B4441AA2 3E889F47 F8881F26 4718228F E8C5D8CA B4419D29
49FAF217 B785C786 C349988A 4E2E1D99 C764712B 875C1B7D 5D825A4A A68C5C4D

8888B4AB 4D2BB694 C24B31DF B6114294 C5C441A2 7A89C197 4B013B22 D78DCF9D
A5319EA3 43CA7C27 D815EBF4 EBF19E58 45836676 B545D469 897199EF 459A5B44

8C9682CA CE5B48E1 3A1E9E9E DDF7593E F5F75B27 F1999EEB 4414F71C CA55F75D
D5819D3B 1D2492DA C211994C 4AFBBFD3 D78D8245 B8319EBD 88E61633 B338A558

CF1EBE54 E199996F 4A8194AB 8985A238 B4F19CEB 7A89C197 7819D183 D472686F
B744559C
```

```
DEBUG COM1 512 PAUSE 18-89-96.LOG 61981 99.5 7 19:11:12
```

Fig. 2. The software can display the raw codes coming off the airwaves in this special debug mode. Seeing the raw codes helps adjust the interface circuit if you're having difficulty reading the signal with the particular radio you're using.

The Software. The PC-compatible software, PD-203.zip, is relatively small at only 90 kilobytes, and can be downloaded directly from the POCSAG Web page. It must be de-archived using PKUNZIP or similar decompression software.

The main program file, PD.EXE, is only 79 kilobytes in size. Written by Peter Baston, GW0PJA and AA2DZ in England, it is distributed as shareware. The trial version will time out after about fifteen minutes of use. At that time, storing the decoded pager data to disk is disabled. The registered version of the software can be ordered from L0ght Heavy Industries (see their POCSAG Web page), or the author.

The decoding software is designed to run under MS-DOS. Because the software does all of the decoding, running it in a DOS box under Windows is not recommended because of the overhead Windows produces. The program is configured by editing the PD.INI file. Serial ports 1, 2, 3, or 4 can be used for message input. The bit rate can be selected manually or be set automatically as messages are received, which will decode all bit rates. Display colors for the background, foreground, and status line can be changed to any combination desired. The program is designed for international use and output can be optimized with spe-

cial characters used in English, German, Swedish, and Danish. The registered version software allows a second serial port to be used to output data in ASCII form.

The PAGERS.INI file is used to specify up to 250 different pager addresses with a seven-digit pager ID code. Whenever a specified address is encountered, it is highlighted on the screen and a beep, if configured, is sounded. Wildcards are allowed in the PAGERS.INI file. The REJECT.INI file is used to specify a list of addresses to be rejected to reduce screen clutter. A typical data stream is shown in Listing 1.

The program can be switched between NORMAL and DEBUG modes by pressing the computer's F1 key. The DEBUG mode, shown in Fig. 2, displays the raw POCSAG codes in hexadecimal format. That is particularly useful for setting up the hardware interface as it gives a visual indication of the number of processing errors in real time. A very helpful status line is shown across the bottom of the screen. That line displays the serial port being used, the current POCSAG bit rate, a PAUSE/RUN indicator, an indication of relative receiving efficiency, a rotating signal indicator, and the current time (taken from the computer's clock). The relative receiving efficiency is expressed in percent. An indication of 100% indicates that all received codes contain no errors. The rotating signal indicator appears to spin when data is being received on the correct pin at the serial port.

The entire project is quite educational. Included with the shareware is a file called POCSAG.TXT written by Brett Miller, N7OLQ, which gives an excellent explanation of the technical aspects of POCSAG signaling and pager operation. It is quite well written, and is perfect for those who are interested in the technical end of pager operations.

In Case Of Difficulty. Many unsuccessful electronics projects can be traced to power supply problems. The first place to check for wiring

continued on page 60

only \$35. There are other combinations of elements and fees for the other elements.

Preparing For An Exam. The best way to prepare for any of these exams is to study the available background subject material. ISCET offers excellent, inexpensive study materials that will help all candidates prepare for each of its exams. If you are at the entry level, the Study Guide for the Associate CET Test will give you an excellent review for this first test. The 96-page booklet is priced at \$10. The Software Study Guide and Practice Test with 300 sample questions is priced at \$39.95 plus \$2 shipping. In addition, ISCET offers practice tests for most of the Journeyman options as well as excellent review texts on each of those options.

The FCC examinations are assembled from questions in a published question pool. By making the complete question pool available, the FCC has defined the limits of the basic knowledge that it expects each successful candidate to have. The availability of the pool also assures all persons taking the test that there will be no nasty surprises. A study guide and compete question pools for Elements 1, 3, and 8 is available for \$29.95 plus \$3 for shipping and handling costs. In addition, self-test computer software packages are available. Not surprisingly, being well prepared for all examinations can make the difference between passing and failing!

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See you at Testing Week 97, and on Electronics Technicians Day! Ω

MOD BOX

continued from page 49

Mod Box is in the Clean mode.

Adjust the Volume control to a comfortable sound level. Strum a few chords and notice that the sound is more or less unaltered. Now crank up the Low control and listen for the full rich boom of the bass notes. Turn it in the opposite direction, and observe how the low end is attenuated. Test the Mid and High controls in a similar manner. By the way, a boosted Mid is great for imitating the cheesy guitars of the early 1960s, while bumping the High gives a great edge to country-western solos.

Switch S1 to distorted, but be sure to watch the Volume control to keep from blowing out your loudspeakers! Dial up the Drive to see how the sound becomes increasingly more ragged. Low settings are perfect for rhythm guitar work, since the sound is quite similar to the creamy distortion of a tube amplifier. For piercing rock solos, spin the Drive control up to its highest setting and notice not only the increase in distortion but the long lasting sustain. And of course, you can further alter the effect by working over the Low, Mid and High controls.

In learning to play any musical instrument, practice is the name of the game. Even though the Mod Box only has 5 potentiometers and 1 switch, there are countless subtle (and not so subtle) effects possible. Experiment with the Mod Box and note the various settings you feel are most useful. After a while, you will probably find the Mod Box to be an indispensable part of your rig. It really can be a natural extension to just about any musical instrument. So what are you waiting for? Build the Mod Box today and see for yourself! Ω

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POCSAG DECODER

continued from page 52

errors is with the op-amp power taken from the RS-232 serial port TxD pin and DTR pin. Pin 7 on the 741 op-amp should read about +10 volts. Pin 4 on the op-amp should read about -10 volts. If either voltage reading is incorrect, check the polarity of D1-D4, C2, and C3.

An easy way to get a visual indication of the operation of the decoder board is with a RS-232 mini-tester, which uses red and green LEDs to indicate voltage polarity. The TxD LED should glow red (for -12 volts), the DTR LED should glow green (for +12 volts), and the CTS (or DSR) pin should flicker red and green to indicate a proper signal output to the computer. Also, as mentioned above,

**TABLE 1—COMMONLY USED
POCSAG FREQUENCIES**

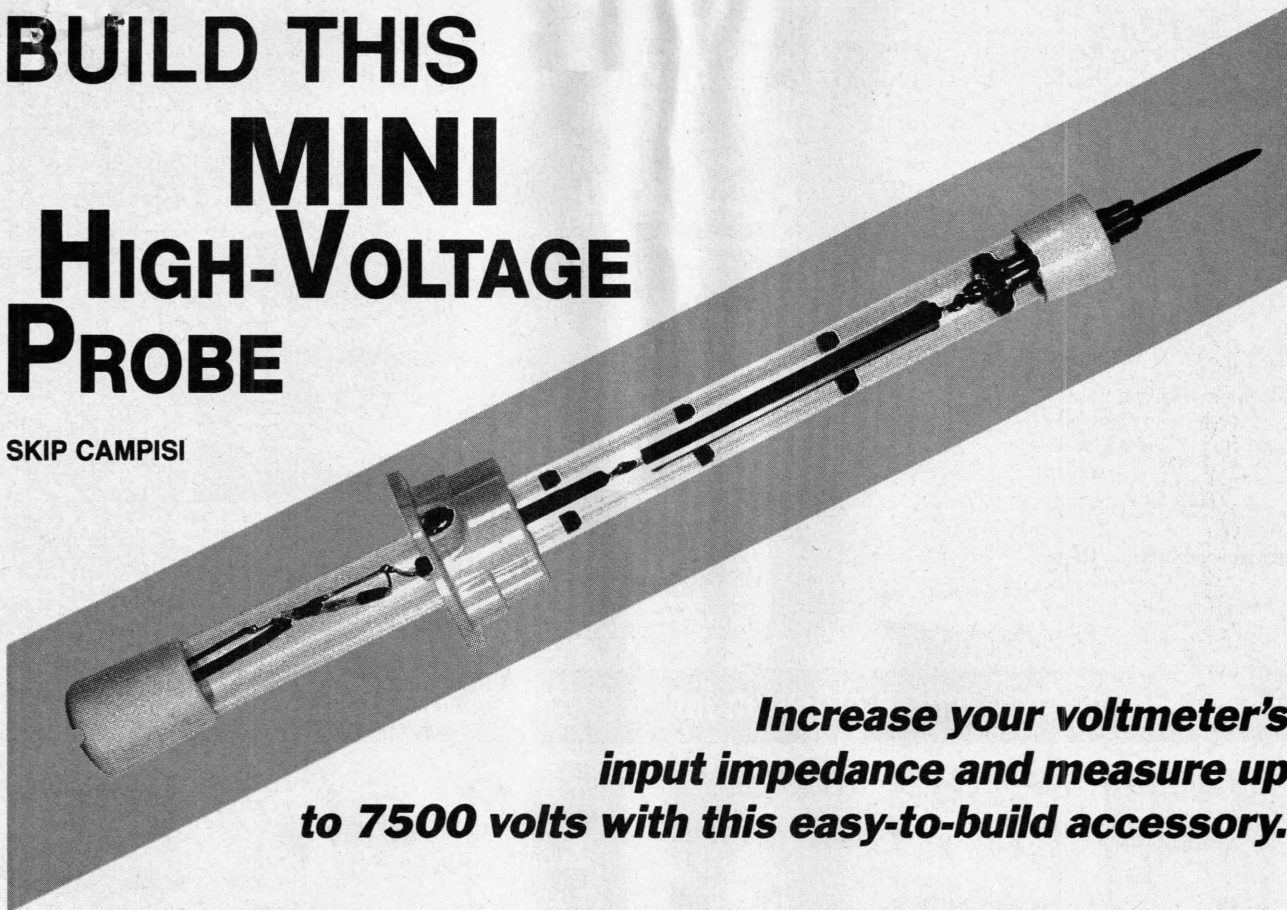
152.03 - 152.24
152.51 - 152.84
158.10
158.70
454.025 - 454.650
931.0125 - 931.0875
931.8875, 931.9125, and 931.9375
are the national channels

the software signal indicator will spin when data is being received on the proper serial port and pin. If a discriminator tap is being used, and everything else appears normal, but the program does not function or many errors are indicated by a low percentage of copy, try either removing R2 from the circuit, or replacing R2 with a jumper. Finally, make sure the software is correctly configured for the proper serial port and the proper data input pin (i.e. either CTS or DSR pin).

Whether you buy the kit or build your own on perfboard, the interface is easy and inexpensive to build. The software is both well written and documented. The project provides an excellent hands-on education with POCSAG signals. But the best reason to try out this project is that it is just plain fun. Ω

BUILD THIS MINI HIGH-VOLTAGE PROBE

SKIP CAMPISI



***Increase your voltmeter's
input impedance and measure up
to 7500 volts with this easy-to-build accessory.***

THE MAXIMUM ALLOWABLE VOLTAGE specification on the typical digital voltmeter (DVM) usually never exceeds 1000 volts. Even the DVM's 10-megohm input impedance every so often loads down a high-impedance circuit giving false readings. That's why you either need a \$100-plus, store-bought, high-voltage probe or the Mini High-Voltage Probe that you can build in one weekend for about \$10 to \$20. So save the money and read on.

The Mini High-Voltage Probe has an input impedance of 110 megohms, or 11 times that of a standard DVM, and it can handle 5000-volts AC or 7500-volts DC peak. The prototype illustrated in this article and a common, service-type DVM were used to test a 5500-volts AC transformer, and the probe performed perfectly. Its accuracy is within 1 percent, which is typical considering the components used in the probe's construction.

About the circuit

The schematic diagram of the

probe (Fig. 1) reveals its simple circuit. It is a standard voltage divider made up of resistors with a 1-percent tolerance. Resistors R1 and R2 are the key elements here. They are rated at 15,000 volts and 10,000 volts, respectively. The 7500-volt DC peak specification for the assembled probe **must not be exceeded under any circumstances!** Play it safe and purchase a factory-built unit if you require higher voltage measurements, especially if you wish to do TV servicing.

The values for series resistors R3 and R4 are in parallel with the 10-megohm resistance of the DVM, thus providing a 1000

to 1 voltage divider. Voltage measurements made by the probe should be multiplied by a factor of 1000. If your DVM uses a different input impedance than the standard 10 megohms, you can adjust the R3-R4 series combination as needed.

Construction

The body of the probe is made from an antenna wall feed-through tube available from Radio Shack. The tube is made of clear plastic with an outside diameter of $\frac{3}{4}$ inch, and a bore of $\frac{5}{8}$ inch. It has two flanged end plates, one of which is glued in position. Remove the other end plate by loosening the Phillips

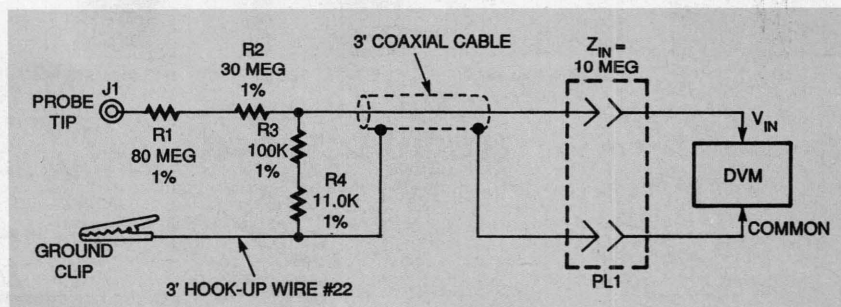


FIG. 1—SCHEMATIC DIAGRAM for the Mini High-Voltage Probe.

head set screw and slide it off the tube. Measure from the open end of the tube and cut off (with a hacksaw or razor saw) an 11-inch length of tube. Discard the short piece with the glued-on end plate.

Take the end plate containing the set screw and note the lip inside the bore. Use a $\frac{1}{4}$ to $\frac{3}{8}$ -inch diameter rat-tail file to smooth down the lip so that the end plate can be slid easily along the tube. This end plate serves as the hilt on your probe and keeps your hand safely away from the high-voltage business end of the probe.

WARNING: Do not be tempted to paint the clear plastic tube—insulation integrity could be compromised!

Now prepare five $\frac{3}{8}$ -inch rubber faucet washers (obtainable at plumbing supply stores or hardware outlets) by enlarging their center bores. The standard $\frac{3}{8}$ -inch washers have an outside diameter of $\frac{5}{8}$ inch with a flat outside edge. Check to be sure that they slide easily into the probe body before drilling. Using drill bits twisted with your fingers (no power tools), you can do a neat and quick job by "stepping up" drill bit sizes in increments of $\frac{1}{64}$ inch or $\frac{1}{32}$ inch to attain the final sizes. Drill two washers to a bore of $\frac{7}{32}$ inch, one washer to $\frac{1}{4}$ inch, and two washers to $\frac{9}{32}$ inch. Install J1, a panel-mount RCA phono jack, in the washer drilled with a $\frac{1}{4}$ -inch bore. Do not over-tighten the nut, or the washer will spread out and bind in the probe's body—a nice slide fit is desirable. Secure the nut on J1 with a dab of silicone sealant.

Drill a $\frac{1}{4}$ -inch hole dead center in the end of a $\frac{3}{4}$ -inch, plastic, chair-leg end cap. Install J2 (panel-mount RCA phono jack) in this hole and tighten securely. Drill an appropriate sized clearance hole in the center of another $\frac{3}{4}$ -inch end cap to pass both the coaxial cable and ground lead you plan to use, allowing about $\frac{1}{16}$ -inch clearance.

Fabricate a probe needle from a 3-inch length of $\frac{1}{8}$ -inch diameter brass or steel rod (the prototype used a section of brass welding rod). Round off one end

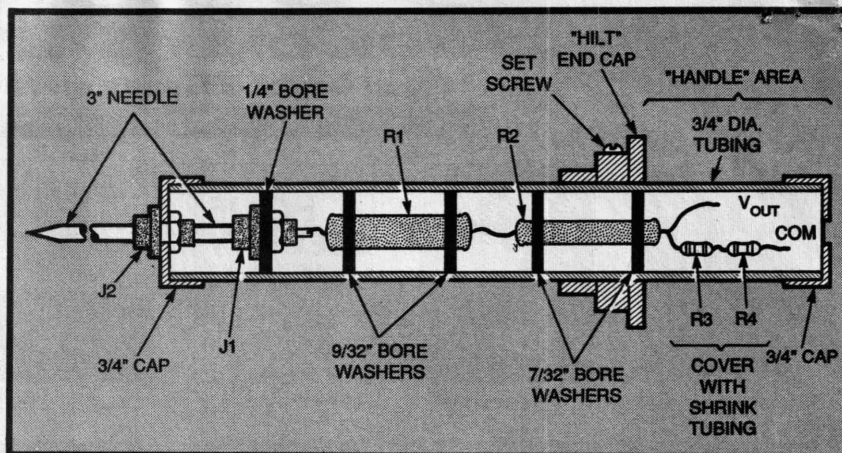


FIG. 2—DETAILED ASSEMBLY VIEW of the Mini High-Voltage Probe. The neoprene faucet washers are identified as $\frac{3}{8}$ -inch; however, their actual outside diameter is $\frac{5}{8}$ inch.

of the rod for easy insertion into the jacks, and file the other end to a blunt needle shape, rather than a sharp point, to suppress arcing and corona discharge. This work can be done by hand, or, more rapidly, chucked in a lathe or power drill. Finish up with fine steel wool and metal polishing paste for a nice finish. Tooth paste works fine in a pinch!

Insert the rounded end of the probe needle into J2, mounted on the $\frac{3}{4}$ -inch cap. Leave about $1\frac{1}{4}$ -inch protruding from the front of J2. The probe needle should be a tight fit to keep it from sliding when used. If the fit is too loose, remove the needle and flow solder into the jack's center conductor pin for a better fit. If necessary, you can solder the needle directly to the pin. Install the assembly of J1 (mounted in its washer) on the rounded end of the needle, leaving J1's center-conductor lug accessible. No connections are to be made to J2's center conductor, or either jack's ground lug.

Refer to the parts location diagram (Fig. 2), then install the two $\frac{9}{32}$ -inch-bore rubber washers on R1, and the two $\frac{7}{32}$ -inch bore rubber washers on R2, in the approximate positions shown. Trim the leads on R1 and R2 to $\frac{1}{2}$ -inch long. Solder R1 to J1's center conductor eyelet. With a long-nose pliers, make wire hooks of the leads on one end of R1 and R2. Connect R2 to R1's other lead by interlocking their wire-hook leads

and solder. Slip the assembly into the probe's body to check for a smooth slide fit. Bend the leads of R1 and R2 carefully to accomplish this.

Series connect R3 and R4, and put a slight offset bend in R2's free lead as shown in Fig. 2. Solder the free end of R3 to R2 near the bend and arrange the leads as shown. Cover R3-R4 with a piece of heat-shrink tubing. Likewise, cover R2's lead with heat-shrink and leave about $\frac{1}{2}$ -inch of exposed wire at the end. Select a 3-foot length of flexible coaxial cable such as RG-174 or audio cable, and solder its center conductor to the junction of R2-R3. Solder the braided shield to R4's free end, along with a 3-foot length of insulated #22 stranded hook-up wire.

Feed the two cables through the probe body followed by the entire resistor/probe tip assembly. If the $\frac{3}{4}$ -inch cap containing the probe tip is not a tight fit on the tube, secure it with a dab of silicone sealant.

Install a cable tie on the two cables right before they exit the tube, and slide a plastic washer up the cables to the tie thus providing strain relief for the cable and wire. Slip the previously completed "hilt" over the cables and onto the tube and position it so that the end of R2's body can just be seen at the junction of R2-R3 (Fig. 2). Tighten the set screw snugly, using care not to crack the tubing. Slip the remaining $\frac{3}{4}$ -inch plastic cap

over the cables and onto the probe's body.

Solder an alligator clip to the end of the 3-foot hook-up wire, which is the probe's ground lead. Select the connector which mates to your DVM (a dual banana plug is typical) and connect it to the end of the 3-foot coaxial cable, the braided shield going to the ground, or common terminal. This completes construction of the probe, which is now ready for testing.

PARTS LIST

Resistors

R1—80 megohms, 1%, 7.5-watt, 15,000-volts, metal-film Caddock #MG780 (Johnson #160-CAD-80M)

R2—30 megohms, 1%, 3.5-watt, 10,000-volts, metal-film Caddock #MG735 (Johnson #160-CAD-30M)

R3—100,000 ohms, 1%, ¼-watt, metal-film (Johnson #151-100K)

R4—11,000 ohms, 1%, ¼-watt, metal-film (Johnson #151-11.0K)

Connectors

J1, J2—Jack, RCA phono, panel mount

PL1—Dual banana plug

Miscellaneous

1—Tube, antenna through-wall lead-in (Radio Shack 15-1200)

2—Cap, plastic chair leg, ¾-in.

5—Washer, neoprene, faucet (sink) ¾-in. (½-in. O.D.)

3-ft.—Coaxial cable, flexible, any impedance

3-ft.—Wire, #22 stranded, insulated, hook-up

1—Alligator clip

All four resistors are available from: Johnson Shop Products, P.O. Box 2843, Cupertino, CA 95015; Tel: 408-257-8614.

Testing and operation

Check for proper operation of the divider circuit on a low-voltage DC source. Plug the probe into your DVM, making sure that the braided shield is connected to the DVM common or ground terminal. Connect the probe ground clip to the negative terminal of a nine-volt battery and touch the probe needle to the positive terminal. The DVM should indicate about 9.0 millivolts as a satisfactory read-

ing. Now measure the battery voltage directly (without the probe) using your DVM and ordinary test leads.

Multiply the probe's reading by 1000, and compare it to the battery voltage direct reading (without the probe). The two readings should match within about $\pm 2\%$ or better. If not, go no further before finding and correcting your problem! When the readings agree, you may check the high-voltage performance of your probe.

WARNING: To avoid the risk of electrical shock and possible damage to the probe or other equipment, follow these instructions exactly!

As a general precaution before proceeding with any high-voltage check with any high-voltage probe, always inspect the insulation integrity of your probe and cables first. Make sure there are good connections to the plug and ground clip. Check that the wire to the ground clip is not worn or frayed. Plug the probe into the DVM, observing proper polarity (ground to ground), and turn on the DVM.

Begin by using a high-voltage AC transformer rated less than 5000-volts AC, with the power OFF. Connect the probe ground clip to one of the transformer's terminals, and by means of a short clip lead, connect the probe needle to the other terminal, keeping the leads separated to avoid drawing an arc. Without touching the probe's body, apply power to the transformer and look and listen for any arcs or "sizzling" at any points in your set-up.

If you have a problem, repair it before attempting to handle the probe during actual use. If everything looks satisfactory, note the actual reading on your DVM, and if this reading appears correct, you are ready to make high-voltage measurements with your home-brew probe!

To measure high-voltage DC, first connect the ground clip to chassis ground, then touch the probe needle to the high voltage terminal. If possible, make the high-voltage connection before

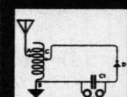
turning on the high-voltage supply: This will prevent arcing to the probe needle, which could damage the probe or the equipment under test. Remember to multiply all of your readings by 1000, and hold the probe by the handle behind the hilt.

Be cautious

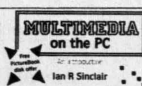
Never grasp the probe ahead of the hilt near the divider section! Never connect the ground clip to a voltage referenced above or below earth ground—serious electrical shock and instrument damage could result. The clip has to be either transformer isolated from ground, or at ground potential for safe operation. Of course the probe can be used for a low-voltage, high-impedance measurements by DVM, also—just remember to use the 1000 multiplication factor. □

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JACOB'S LADDER

Build this exciting Jacobs's Ladder and watch electric arcs ascend the ladder and evaporate in space. It works from a clever 12,000-volt power supply.

ROBERT IANNINI

PEOPLE HAVE LONG BEEN FASCINATED BY ELECTRIC arcs—and perhaps put off by them. They show up as lightning, Tesla coil discharges, and long sparks that sting as you reach for the doorknob on a cold, dry, winter day. This Jacob's Ladder project turns electric arcs into a dramatic but harmless conversation piece.

If you build this project, you'll learn how a simple power supply operating from the 120-volt AC line can produce 12,000 volts. In addition to powering the Jacob's Ladder, the supply can power plasma displays, and it has even powered a light-duty, bench-type spot welder.

Perhaps you would like to know the origin of the term Jacob's Ladder. The Bible tells the story of Jacob's dream about a ladder that extended from earth to heaven. Jacob, the son of Isaac, was the father of the founders of the twelve tribes of Israel. Among sailors, however, a Jacob's Ladder is a long rope ladder that is hung over the side of a ship so the harbor pilot can climb aboard.

Climbing arcs

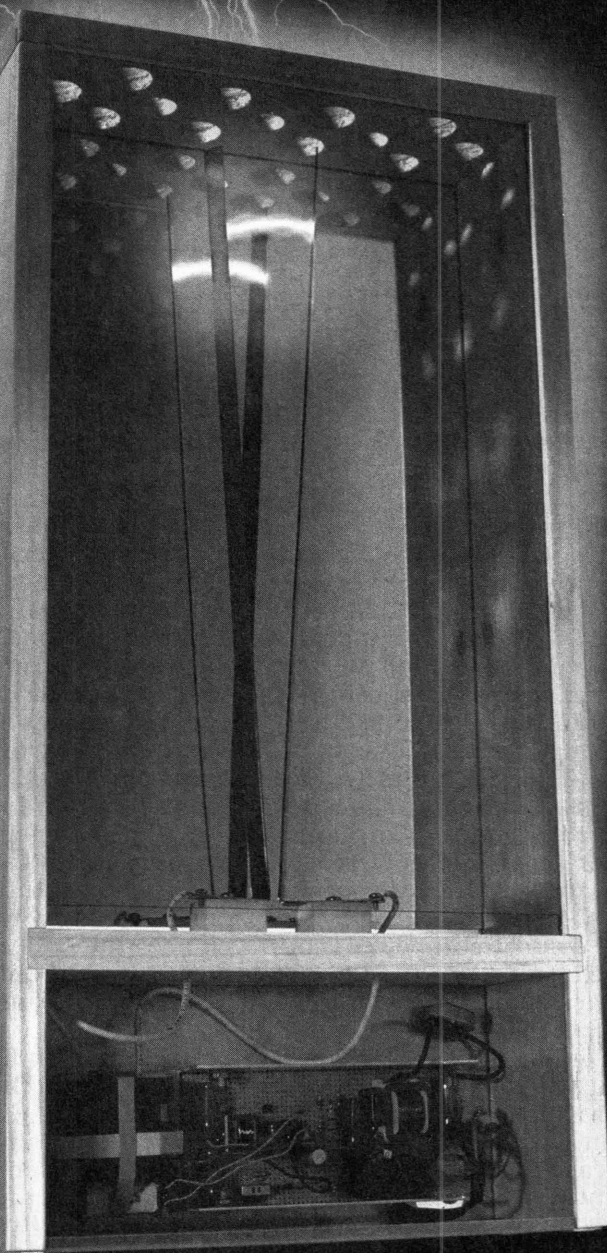
The power supply for this project forms electric arcs across two diverging stainless steel strips mounted in a protective case. The 16-inch long strips are mounted on insulating Teflon blocks to eliminate possible leakage. The stainless steel strips are angled with respect to each other so that the arcs form at the edges of the strips that are separated by about $\frac{3}{16}$ inch at their bases, but the strips diverge to a distance of about 2 inches at their upper ends.

The strips form a gap in the secondary winding of the output transformer. After power is turned on, the air dielectric breaks down due to the "almost" short-circuit state across the lower end of the gap, and an electric arc is formed.

As the arc heats up, thermal convection causes the arc to rise up the vee-shaped "ladder." As the plasma arc ascends the ladder, its length increases, thereby increasing the arc's dynamic resistance and thus increasing power consumption and heat. This causes the arc to stretch as it rises and extinguish when it reaches the top of the ladder. When the arc extinguishes, the transformer output momentarily exists in an open circuit state until the breakdown of the air di-

electric produces another arc at the base of the ladder and the sequence repeats.

The power supply for the Jacob's Ladder contains circuitry to protect persons and property from electrical shock or fire hazard if the ladder strips should be shorted accidentally when the ladder is operating.



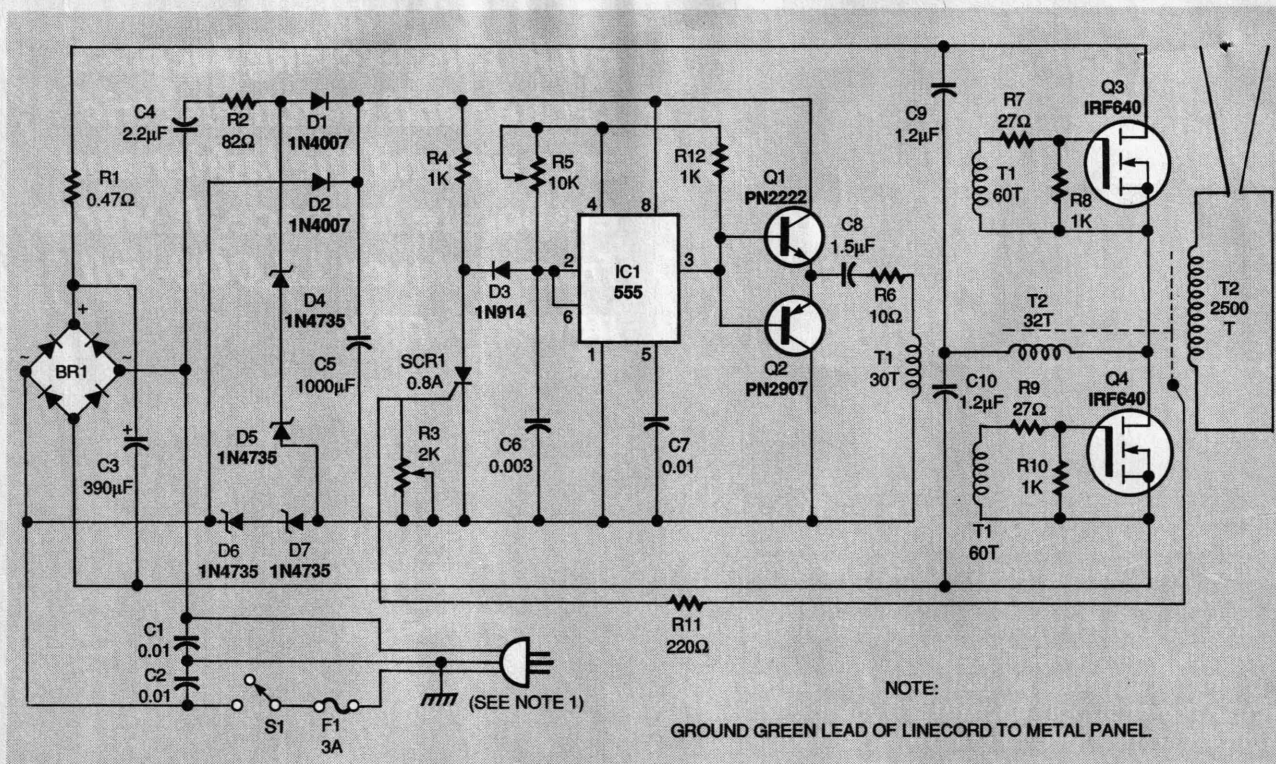


FIG. 1—SCHEMATIC FOR HIGH-VOLTAGE power supply. All components except transformers are standard.

12,000 volt supply

The operation of the Jacobs's Ladder depends on the current-limited power supply that delivers 12,000 volts at 40 milliamperes from the 120-volt AC line. Refer to the schematic Fig. 1. The full-wave bridge BR1 rectifies the 120-volt AC line input, and resistor R1 limits the DC charging current of capacitor C3 to a safe value. The Jacob's Ladder project is powered by 150 to 160 volts DC derived from the 120-volt AC line.

The drive circuit power is obtained by dropping the 120-volt AC line through capacitor C4 and current limiting resistor R2. Diodes D1 and D2 alternately conduct only in the positive direction, and their DC output pulses are integrated by filter capacitor C5. Zener diodes D4 and D5 regulate these DC pulses to peak values of 15 volts DC.

Capacitor C4 and resistor R2 present a complex impedance so that most of the AC line voltage is dropped across the reactance. This configuration eliminates the waste of real power and heat losses that would have occurred by dropping the line

voltage with only a resistor.

The 555 timer IC1 is configured as a square-wave oscillator. The output frequency of its square waves is determined by the setting of trimmer potentiometer R5 and capacitor C6. The frequency is about 25 kHz for the values of R5 and C6 shown in Fig. 1. Resistor R12 limits the high-frequency setting to an acceptable value. Potentiometer R5 can be used to adjust the circuit's power output. Increasing the frequency reduces the circuit's output by increasing the inductive reactance of the transformer leakage inductance.

The output of IC1 on OUTPUT pin 3 appears at the bases of the current "source," NPN transistor Q1 and "sink" PNP transistor Q2. The emitters of this transistor pair are AC-coupled through capacitor C8 and resistor R6 to drive the primary of driver-isolation transformer T1. This drive prevents DC from flowing through the primary. Resistor R6 dampens any overshoot that results from transformer T1's leakage inductance.

This scheme provides a satisfactory source for driving the

high gate-to-source capacitance of the two NPN switching power MOSFETs, Q3 and Q4. Transformer T1 is wound on a high-permeability core with as few turns as possible to eliminate leakage inductance.

The gate circuits of MOSFETs Q3 and Q4 contain 27-ohm resistors (R7 and R9) to slow their switching times. This eliminates possible parasitic oscillations that could occur if the MOSFETs were switched at their speed limit.

Output transformer T2 has a half-bridge configuration so that MOSFETs Q3 and Q4 are only subjected to half of the rectified DC line voltage, or about 80 volts. Assuming that the voltage midway between capacitors C9 and C10 is about half of 160 volts (or about 80 volts), when Q3 turns on, the charge on C9 causes current to flow through T2 in one direction.

When Q3 turns off, Q4 turns on, dumping the charge on C10 through T2 in the opposite direction. This drives the magnetic flux of T2 evenly and symmetrically, making full use of T2's core capability. The primary of output transformer T2 contains 32 turns, but its sec-

Continued on page 56

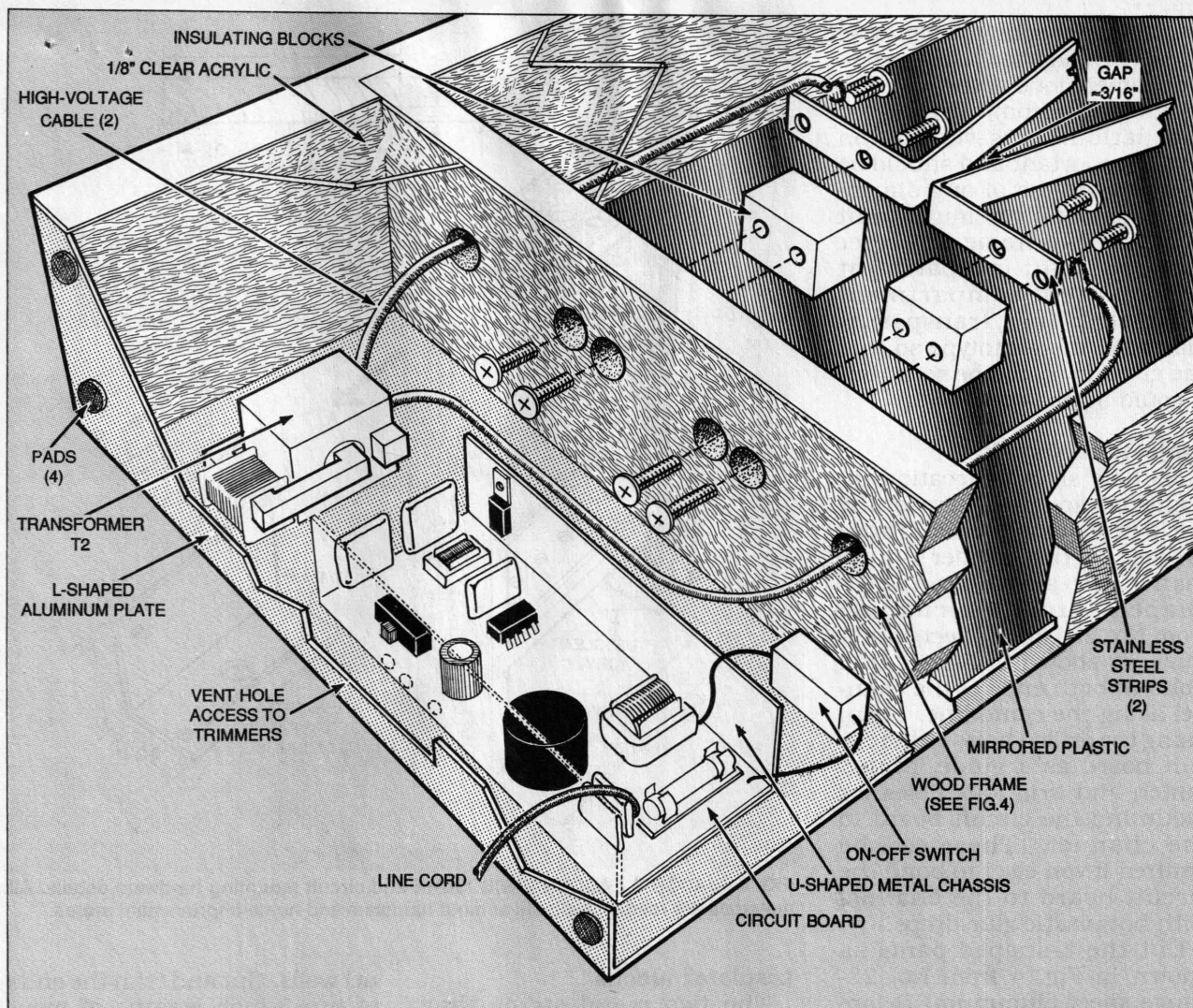


FIG. 3—ASSEMBLY DIAGRAM FOR JACOB'S LADDER. All circuitry is enclosed with insulating wood and protective plastic covers.

quirements by providing adequate insulation between people and flammable materials and the enclosed high-voltage circuitry, while at the same time protecting the circuitry from dust and dirt. 2. to be simple enough to be made by persons with minimal carpentry skills from materials readily available at hardware and home-improvement stores.

Many variations on the author's enclosure design are possible including changes in exterior and interior dimensions and the substitution of more expensive wood for the framing. However, it is imperative that all provisions for ventilating both the circuitry and the vee "ladder" to dissipate any heat buildup be followed, but

make sure that those spaces are not large enough to admit fingers or the small hands of curious children.

The overall case dimensions are $24 \times 12\frac{1}{2} \times 4$ inches. The closed H-shaped frame was made from $\frac{3}{4}$ -inch thick $\times 4$ -inch wide soft wood. Slots that are $\frac{1}{8}$ -inch wide and $\frac{1}{4}$ deep were milled $\frac{1}{8}$ -inch in from each edge of the inside surfaces of the frame members to accommodate protective transparent plastic covers on the front side and a metallized plastic mirror on the back side. (These protective sheets could be fastened directly to the case edges with screws.)

The bottom of the case and the back of the lower circuit/transformer compartment is

covered by an L-shaped aluminum plate that serves as the vertical support for the circuit board and output transformer. Holes drilled in the bottom of this plate permit circuit ventilation and access to the on-board trimmer potentiometers R3 and R5. Another hole is formed in the back of the plate for mounting ON-OFF pull switch S1.

Figure 4 provides general information on the sizes and shapes of the principal wood and aluminum parts. Notice the holes drilled in the top member of the frame for cooling the ladder compartment. The author's prototype frame was made by assembling the wooden frame with screws after the clear $\frac{1}{8}$ -inch plastic front windows and rear mirror were cut to fit the milled slots.

The transparent plastic cover

for the ladder compartment was cut 1-inch shorter than the inside dimensions of the frame to provide a bottom opening for ventilation. This ventilation slot is important and should be there regardless of any dimensional changes you might want to make in the frame. The cover for the circuit compartment protects that compartment completely. It is transparent plastic in the prototype so that the circuitry could be seen, but it could be opaque.

Project metalwork

No hole sizes or location dimensions are given here for the enclosure components. Those are left to the builder's judgment. Cut and form the U-shaped aluminum channel from No. 22 gauge sheet aluminum, as shown in Fig. 4. Drill holes in both ends of the channel along the centerline. Then, using the drilled holes in the circuit board as a guide, center-punch and drill four holes for mounting the circuit board to the channel. (These can be omitted if you elect to bond the circuit board to the channel with hot plastic glue drops.)

Cut the L-shaped panel as shown in Fig. 4 from No. 22-gauge sheet aluminum. Before bending the front lip or folding the plate, drill 1-inch diameter holes for circuit cooling and access to trimmer potentiometers R3 and R5. Above the fold line, drill another hole for the linecord, large enough to admit a rubber or plastic grommet, and drill a hole to accommodate chain-pull switch S1. Then bend the flat plate 90° along the fold line and bend up the front lip.

Drill two holes evenly spaced within 2 inches of the ends of the stainless steel "ladder" strips, and bend those 2-inch long sections approximately 90° with respect to the rest of the strips. Note: Stainless steel was selected for the ladder strips because the electric arcs will not cause the strips to oxidize or corrode, and tests showed that stainless steel permits easier starting of the arcs than other metals.

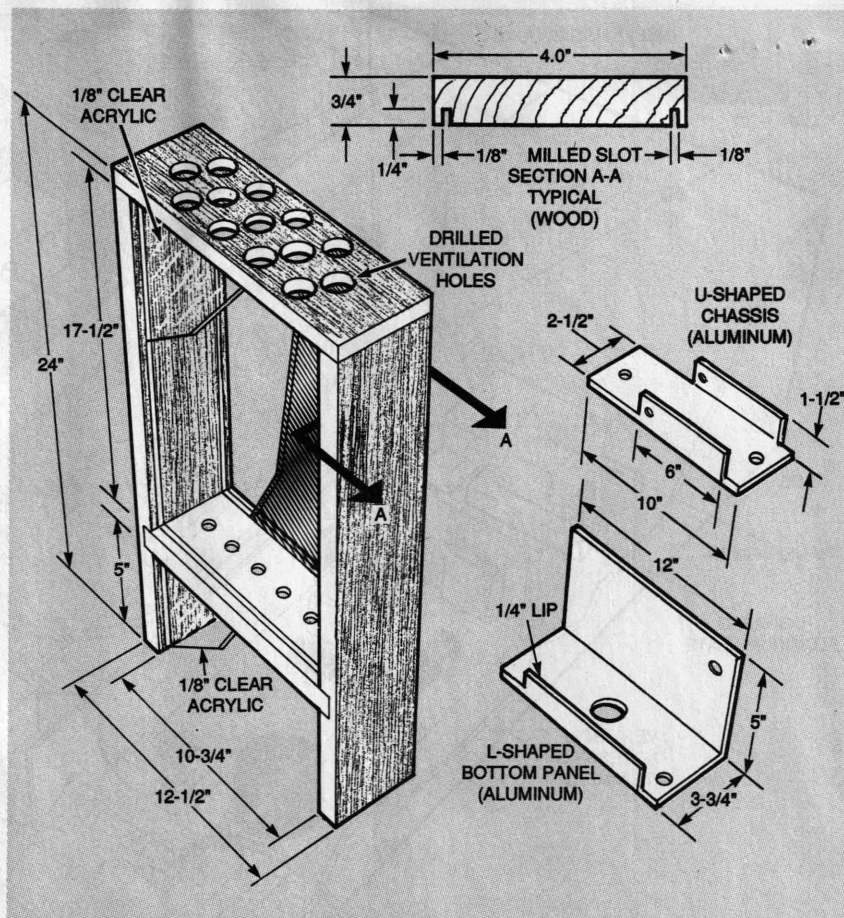


FIG. 4—JACOB'S LADDER ENCLOSURE and circuit mounting hardware details. All materials are readily available at most hardware and home-improvement stores.

Insulator blocks

The two metal strips that form the "ladder" must be mounted on insulators that have high dielectric strength. The insulators in the prototype were made from Teflon blocks that measure $1\frac{1}{4} \times 1 \times \frac{3}{4}$ -inch high. This material can be drilled and tapped, and it is strong enough to withstand the heat created by electric arcs. Individual Teflon blocks are available from the source given in the parts list.

Project assembly

With the completed circuit board inserted in the channel, elevated slightly above the bottom of the channel, mark, center-punch and drill the holes in each side wall for fastening the tabs of MOSFETs Q3 and Q4. Be sure to deburr and perhaps countersink slightly the holes in the channel so that the tabs on the MOSFETs will be clamped securely against the chan-

nel walls. Cut and trim the ends of two 3-inch lengths of insulated, stranded linecord to the circuit board, as shown in Fig. 2, to make the connections with the linecord. Insert and solder the ends of the wires on on-off switch S1.

Attach the circuit board to the U-channel with screws and nuts, using several nuts as standoffs to isolate the circuit board from the channel. Align the tabs of MOSFETs Q3 and Q4 with the holes in the sidewalls of the channel, insert insulating mica washers with a film of silicone grease between each tab and the channel walls, and fasten them with screws and nuts.

Cut a 2½-inch square of phenolic laminate or circuit board stock and place it between the base of transformer T2 and the bottom of the channel. Bond the transformer to the insulator and channel base with epoxy or hot glue.

PARTS LIST

All resistors are 1/4, 10%, unless otherwise specified.

R1—0.47 ohm, 2 watt

R2—82 ohms, 2 watt

R3—2000 ohms, PCB-mount, trimmer potentiometer

R4, R8, R10, R12—1000 ohms

R5—10,000

R6—10 ohms, 1/4 watt

R7, R9—27 ohms

R11—1229 ohms

Capacitors

C1, C2—0.01 μ F, 1000 volts, ceramic disk

C3—390 μ F, 200 volts, aluminum electrolytic

C4—2.2 μ F, 250 volts, metallized polyester

C5—1000 μ F, 25 volts, aluminum electrolytic

C6—0.003 μ F, 50 volts, polyester film

C7—0.01 μ F, 50 volts, ceramic disk

C8—1.5 μ F, 100 volts, metallized polyester film

C9, C10—1.2 μ F, 400 volt, metallized polyester film

Semiconductors

BR1—bridge rectifier, silicon diode, 4 ampere, SIP

D1, D2—1N4007, 1000 volts, 1, ampere, silicon diode

D3—1N914, silicon diode

D4, D5, D6, D7—1N4735 Zener diode, 6.2-volt, 1 watt

SCR1—silicon controlled rectifier 0.8 ampere, 200 volt, sensitive gate, TO-92, Teccor EC103B1 or equiv.

Q1—2N2222, NPN transistor

Q2—2N2907, PNP transistor

Q3, Q4—IRF640 N-channel power MOSFET, 200 volt, 10 ampere, International rectifier or equiv.

IC1—555 timer

Magnetics

T1—driver transformer, 30 turns primary, 60 turns two secondaries (see

text)

T2—output transformer, half bridge, 32 turns primary, 2500 turns secondary (see text)

Other components

S1—switch, SPST, panel-mount, 10 ampere, pull chain

F1—fuse, 3 ampere, slow-blow

Miscellaneous: uit board or perforated board (see text); U-chassis, (see text); mirror, metallized plastic (see text); transparent plastic covers (see text); L-panel, (see text) stainless steel strips, 0.060-inch thick, 18-1/2 inches (see text) insulating blocks (see text); high-voltage wire, 15 kilvolt rating, 2 feet; 3-wire power cord with line plug; fuse holder; phenolic transformer insulator (see text); hookup wire; miscellaneous nuts, screws; rubber feet, four; linecord grommet; silicone grease; TO-220 mica washers; solder.

Note: The following parts and kits are available from Information Unlimited, Box 716, Amherst, N.H., 03031; Telephone 603-673-4730; Fax 603-672-5405:

- Kit of electronics components and circuit board, less transformers—\$69.50

- Assembled and tested electronic circuit board—\$89.50

- Assembled and tested driver transformer T1—\$9.50

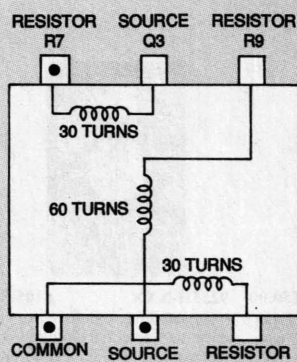
- Cores and bobbins for driver transformer T1—\$4.50

- Assembled and tested high-voltage transformer T2—\$24.50

- Cores, bobbins, and potting cap for high-voltage transformer T2—\$14.50

- High-voltage wire (15 KV)—\$0.50/foot

- Teflon insulating blocks (2 required)—\$5.00



NOTES:
DOTS ON PINS INDICATE STARTING POINTS FOR ALL WINDINGS.
FOR SOURCE OF CORES AND BOBBIN, SEE TEXT.

FIG. 5—TRANSFORMER T1 WINDING. This is the "footprint" for driver transformer T1.

Drill two holes through the wooden frame member between the two compartments, and drill mating holes in the two insulating blocks. Drill a single hole in the opposite sides of each block and fasten the ladder strips to the blocks with screws, as shown in Fig. 3.

Then attach the insulation blocks and strips to the frame member with screws. Fasten the bent metal strips so that they are offset by about 30°, as shown in Fig. 3. The base strips are diagonal across the insulator blocks so that the corner edges are separated by about 3/16 inch. The upper ends of the strips should be about 2 inches apart.

Fasten the channel with

transformer T2 and the circuit board to the L-shaped panel with screws and nuts, as shown in Fig. 3. Complete the installation of switch S1 and the linecord with a grommet, and complete all soldering. Be sure the L-shaped metal panel is hard-wire connected to the earth ground by means of the green wire within the three-wire power cord. Assemble the enclosure with its plastic covers and aluminum base plate. The Jacob's Ladder is now complete.

Adjusting the ladder

Carefully examine your work to be sure that there are no inadvertent short circuits or cold solder joints. The circuit is now ready for testing. First, adjust potentiometer R5 as follows:

1. Disconnect one end of R1 to Q3.
2. Connect an oscilloscope to the Q4 gate.
3. Plug in power cord and turn on power.
4. Adjust R5 for a period of 40ns so 15-volt squarewaves are symmetrical.
5. Shut off power, reconnect R1, and connect the oscilloscope to the Q4 drain. Turn on power. A near perfect squarewave should appear, and it should remain constant as arcs form and reform.

Adjust the ground-fault potentiometer R3 to shut off the circuit if there is ground-fault current.

1. Turn off power and set wiper of R3 for maximum sensitivity.
2. Disconnect a high-voltage lead from a ladder strip, and connect it to ten 1-watt, 100 kilohm resistors in series. (This simulates a 10 milliamperere ground current.)
3. Apply power and verify that the circuit shuts down. Turn off power, adjust R3 slightly clockwise. Reapply power until high voltage stays on. Note: Allow time for C3 to discharge before reapplying power or SCR1 will stay on.

Winding the transformers

Figure 5 is a "footprint" diagram of drive transformer T1 with callouts that can be related

Continued on page 87

EDWARD BARROW

THE REGULATED POWER SUPPLY CIRCUIT described in this article will permit you to experiment with electrolysis and electroplating. It is intended for the experimenter or hobbyist who wants to perform experiments in electrochemistry and electroplate small objects—tools, rings, spoons, or miscellaneous hardware items.

This power supply will also be especially valuable for science and chemistry teachers or laboratory assistants in high schools, technical schools, and colleges. It is a versatile power supply that can be controlled to adapt it to various demonstrations or experiments in chemistry, physics, biology, or general science.

This power supply permits better control of its output voltage than a typical linear power supply, and it can function either as a constant-voltage or constant-current source. Both functions can be set by a potentiometer with a voltage range of 0 to 10 volts and a current range of 0 to 1.0 amperes. Figure 1-a is a block diagram of the basic regulator circuit.

A constant-current mode can be set to control the current for electroplating. This control is critical for achieving uniform deposition of metal. This mode also permits the user to compensate for electrolyte loss that occurs during electroplating. Figure 1-b is the block diagram for the measurement circuit for the power supply. It monitors the output allowing either the voltage or current value to be read on a two-digit LED digital display.

This display is especially useful in electroplating because the voltage must be monitored continuously. Applied voltage determines the reactions that will take place. If

the voltage is too high, the quality of the plating will suffer.

The digital display can also record the total amount of charge passed from the output. In electrochemistry, this quantity is directly related to the extent of ionization that has occurred within the electrolyte. In electroplating, the thickness of the plating deposited is directly proportional to the charge in coulombs passed per second.

Electrochemical theory

Before building this power supply, you might want to review the fundamentals of electrochemistry. This subject is covered in most introductory high school chemistry text and college level inorganic or general chemistry texts. There are

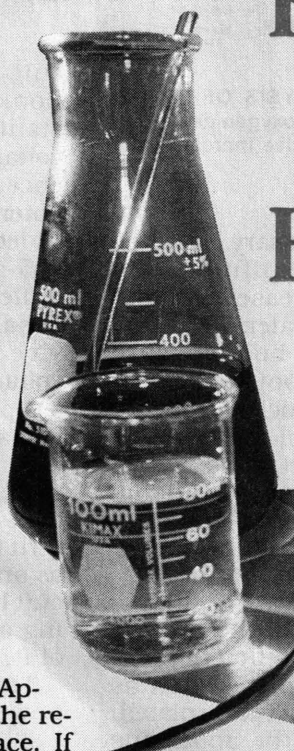
also self-teaching texts in general chemistry keyed to do-it-yourself experimentation available.

Even if you have never taken a formal course of instruction in general chemistry, you can learn the fundamentals of electrolysis and electroplating by studying one or more of the available texts. Some of the important terms in electrochemistry that you should know are defined in the box. *Common Electrochemical Terms Defined.*

In electrochemistry, reactions involve only electrons (hole motion is not discussed). Elements bond either by sharing (covalent bonding) or transferring electrons from one another (ionic bonding). Electrochemistry is concerned only with ionic bonding. It is possible to reverse

Build this regulated electrochemistry power supply and learn about electrolysis and electroplating. Then plate small objects to protect them and improve their appearance.

Regulated Power for Electrochemistry



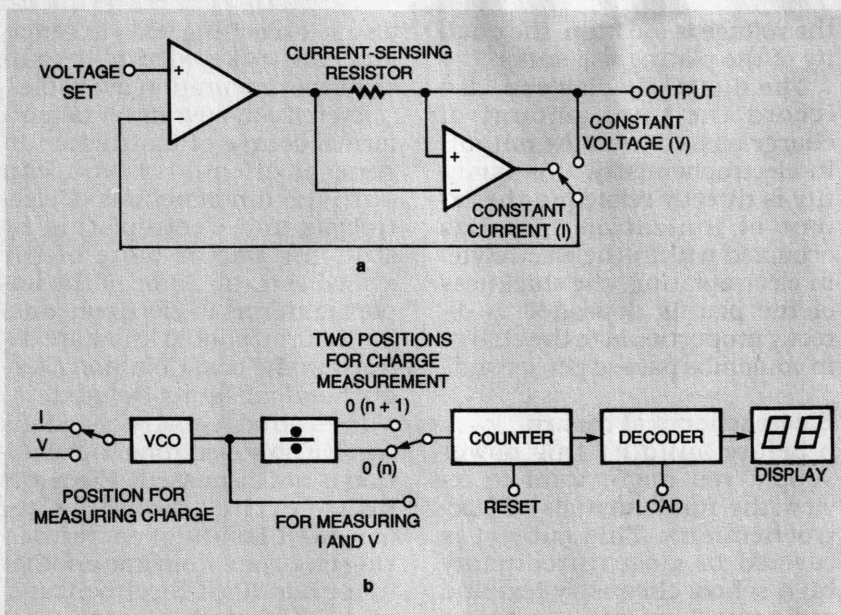


FIG. 1—BLOCK DIAGRAM OF REGULATED POWER SUPPLY CIRCUIT: Basic regulator circuit (a), and measurement circuit (b). Current and voltage or charge can be measured. Values can be read on the LED digital display.

the electron transfer process with current.

Electrolysis explained

Nearly 200 years ago it was found that water breaks down into hydrogen and oxygen when an electric current is passed through it. This was discovered after the Italian scientist, Alessandro Volta, made the first electric battery capable of producing continuous current. When wires were connected to both terminals of a battery and their ends were dipped into a tank of water to act as electrodes, it was found that gas bubbles formed in the water at each electrode.

When the gas was collected, it was found that hydrogen gas (H_2) was given off at the negative electrode and oxygen gas (O_2) was given off at the positive electrode. The volume of hydrogen collected was twice the volume of oxygen in the same time the apparatus was operating. It was then discovered that some of water had been decomposed into its elements. This process was called *electrolysis*.

The formula for electrolysis of water is:

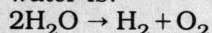


Figure 2 shows a glass Hoffman laboratory apparatus organized for the electrolysis of

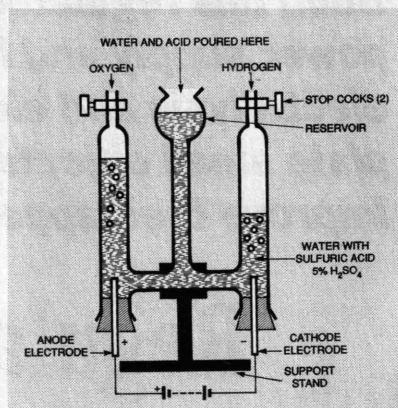


FIG. 2—ELECTROLYSIS OF WATER yields hydrogen and oxygen gas as by-products. Acid additive increases the conductivity of water.

water. It is necessary to add 1 part in 20 of sulfuric acid (H_2SO_4) to increase the conductivity of the water. While this experiment can be done with separate small bottles or test tubes, the stop cocks at the top of each column of the Hoffman apparatus permit the gases to be bled off conveniently and ignited by a burning splint to identify them.

Electroplating.

Electroplating is the process of causing electric current to deposit a layer of metal such as copper on an object to be plated. Figure 4 shows the apparatus

for electroplating. The plating solution is a salt of the metal. In Fig. 4, the electrolyte is a solution of copper sulfate.

A bar or strip of the metal is connected as the anode (positive electrode) to provide positive ions for electroplating. The object to be plated is connected as the cathode (negative electrode) and receives the positive ions that it reduces to a metallic layer of electroplate.

Regulated power supply

Figure 5 is the schematic for the regulated power supply. Its output current is provided by a voltage regulator consisting of operational amplifier IC5-a and transistors Q1 and Q2. The cir-

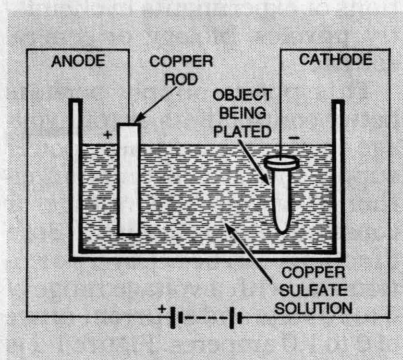


FIG. 3—APPARATUS FOR ELECTROPLATING: The copper strip is the anode and the object to be plated with copper is the cathode.

cuit has two modes of operation: constant-current and constant-voltage. In the constant-voltage mode, a variable reference voltage is generated across potentiometer R34. The voltage is fed to operational amplifier IC5-a, which functions as a buffer. This operational amplifier also serves as a feedback device to monitor the output voltage.

Operational amplifier IC5-a has an inverting input, and its output voltage is altered to match this input with the reference voltage. IC5-a's output is sent to transistor Q1 configured as an emitter follower. This circuit buffers the output, producing a lower impedance version of it.

The output of Q1 is fed to power transistor Q2 which

Continued on page 64

lowers the impedance so that it can source currents up to 1 ampere. The feedback loop consists of resistors R2, R4, R3, and R5. Operational amplifier IC5-a is configured in the feedback loop to compensate for voltage drops across the transistors (i.e., 0.7 volts each). Consequently, the output is a low impedance "mirror" of the input.

A method for measuring the output current and converting it to a voltage allows the circuitry to function as a constant-current source. This voltage can replace the feedback voltage. Then the feedback loop will adjust the output voltage so that this feedback voltage (which represents the output current) is the same as the reference voltage taken from the potentiometer R34. The current will also be set by R34, and the output will behave like a constant-current source.

The voltage across R2 is proportional to the current flowing through it. The 1-ohm value of R2 was selected for connection in series with the output to act as a sensor. The voltage across the resistor is referenced to the output voltage, not the ground. Thus, differential amplifier IC5-b corrects this condition. It responds to the difference in voltage across the resistor and amplifies that difference by a factor of K.

By carefully selecting the values of R and gain or K, a scale is obtained whose current I can be measured in volts. Resistor R and gain K are set so a current of 1 ampere produces 10-volts. Consequently the output current is $V_{REF}/10$.

Measurement method

Electrical charge is current integrated with respect to time. Therefore, to measure the charge passed by the output, the voltage, a measure of the current, is integrated. It will be the output of the differential amplifier IC5-b.

Unfortunately, an analog inte-

All resistors are 1/4-watt, 10% unless otherwise stated.

- R1—1200 ohms
- R2—1.0 ohm, wirewound, 1 watt
- R3—820 ohms
- R4, R6, R14, R15—10,000 ohms, 1%
- R5, R7—100,000 ohms, 1%
- R8, R10—1000 ohms
- R9—33,000 ohms
- R11—68,000 ohms
- R12—2,000,000 ohms
- R13—47,000 ohms
- R16 to R29—3000 ohms
- R30—3900 ohms
- R31—10,000,000
- R32—86,000 ohms
- R33—panel potentiometer, 10,000 ohms, linear taper, 1/2 watt, Mouser 31VC401 or equiv.
- R34—panel potentiometer, 25,000 ohms, linear taper, 1/2 watt, Mouser 31VC403 or equiv.
- R35—trimmer potentiometer, 10,000 ohms, vertical mount

Capacitors

- C1—2.2μF, tantalum electrolytic
- C2—0.1μF, polyester film
- C3—0.1μF, polyester film
- C4—0.3μF, polyester film
- C5, C6—22pF, ceramic (see text)
- C7, C8—0.010μF, polyester film
- C9—10μF, tantalum electrolytic
- C10—4700μF, aluminum electrolytic
- C11, C12, C13—0.010μF, polyester film

Semiconductors

- D1—1N5239B, Zener diode, 9.1 volts, 500 mW
- D2—1N5237B, Zener diode, 8.2 volts, 500 mW
- BR1—, bridge rectifier, 1.5 ampere, 200 PIV, General Instrument WO2 or equiv.
- Q1—2N3705, NPN transistor, TO-92
- Q2—MJE3055T NPN power transistor, Motorola or equiv.

PARTS LIST

- IC1—CD4521B CMOS 24-stage frequency divider, Harris or equiv.
- IC2—CD4011UB CMOS quad 2-input NAND gate, Harris or equiv.
- IC3—CD4046B CMOS phase-locked loop, Harris or equiv.
- IC4, IC6, IC7—CD 4518B CMOS dual BCD up-counter, Harris or equiv.
- IC5—LM358 dual operational amplifier, National or equiv.
- IC8, IC9—CD4543B CMOS BCD to seven segment decoder driver, Harris or equiv.
- IC10—MC7812 three-terminal positive voltage regulator, Motorola or equiv.
- DISP 1—MAN6740, light-emitting diode display, 0.56-inch dual digit, common cathode, RHDP, red, 7-segment or equiv.
- Other components**
- S1, S3—toggle switch, miniature SPDT panel-mount
- S2—rotary switch 3-pole, 12-station
- S4—pushbutton switch, nonlatching, push-to-make
- XTAL1—crystal, 4.194304 MHz, 2-pin metal case
- TR1—transformer 120 volts to 6 volts, panel-mount
- J1, J2—banana plug jacks

Miscellaneous: main circuit board; dual-digit LED circuit board; metal project case, two part (see text); heatsink (see text); knobs for two potentiometers and one switch; No. 18 and No. 22 AWG stranded, insulated hookup wire; No. 28 AWG flat ribbon cable; 120-volt AC linecord with plug; linecord grommet; red plastic filter for LED display DISP1 (1-14 × 1 1/2 inches—see text); assorted nuts, lockwashers and screws; four adhesive rubber footpads, solder, silicone grease.

grator formed from an operational amplifier and a feedback resistor will not work satisfactorily because of the duration of electrolysis or electroplating (over an hour). A very large and expensive capacitor would be needed to meet this requirement, but it would introduce drift and leakage. Moreover, the output would be an analog value that would have to be converted to obtain a digital readout. For this reason, a voltage-controlled oscillator (VCO)

was selected as the integrator.

As a result of including the VCO, the following statements can be made:

1. Output frequency = gain × V_{IN}
2. The number of pulses = frequency × time
3. It then follows that if $V_{IN} = I$
4. Then the number of pulses = $K \times I \times \text{time}$

If gain K is chosen correctly, the number of output pulses effectively equals current integrated with respect to time.

"Sniff" Out Transmitters with the RF Informant

Do more than search for transmitters with this pocket-sized RF-strength meter.

RICK DUKER

Mention the phrase "transmitter detector" and the first image that comes to mind is probably one of searching for hidden surveillance microphones, or "bugs". However, there is a use for such "bug sniffers" beyond the scope of the standard "James Bond" scenario. Examples of using an RF detector on the test bench include checking the operation of devices such as two-way transceivers, ham radios, cellular and cordless telephones, and baby-room monitors. Yet another some-

what unique and interesting application for this device is checking microwave ovens for leakage.

The pocket-sized RF Informant presented here is just such a device. Technically, it can be described as an AM/FM near-field radio-frequency receiver. The relative field strength of any RF signal that is in close proximity to the RF Informant's antenna is monitored, with the field strength displayed on a row of nine LEDs. An audio output is provided for earphone monitoring of the received signals.

The RF Informant operates in two reception modes — *wideband* and *high-band*. In the wideband mode, the receiver is un-tuned and will detect practically any RF frequency from the low AM

band below 500 kHz into the microwave range above 2 GHz. In the highband mode, the receiver is optimized for operation in the FM broadcast band.

And yes, the RF Informant can locate hidden transmitter surveillance "bugs".

Circuit Description. The schematic diagram for the RF Informant is shown in Fig. 1. RF signals arriving at ANT1 are coupled by C5 to the detection circuit. A high-impedance ground connection for wideband reception is provided by R3. With LI switched into the circuit by S2, the circuit is optimized for the FM band.

Diodes D1 and D2 do detection and demodulation. The detected signal is fed to the non-inverting input of IC1. That op-amp is configured as a non-inverting amplifier with a fixed gain of about 450. The particular device specified uses junction field-effect transistors (JFETs) on the inputs; that increases sensitivity due to their high impedance. Potentiometer R9 is a squelch control that adjusts the offset of IC1. The amplified detector output that appears on pin 6 of IC1 is fed to J1. A suitable high-impedance ear-

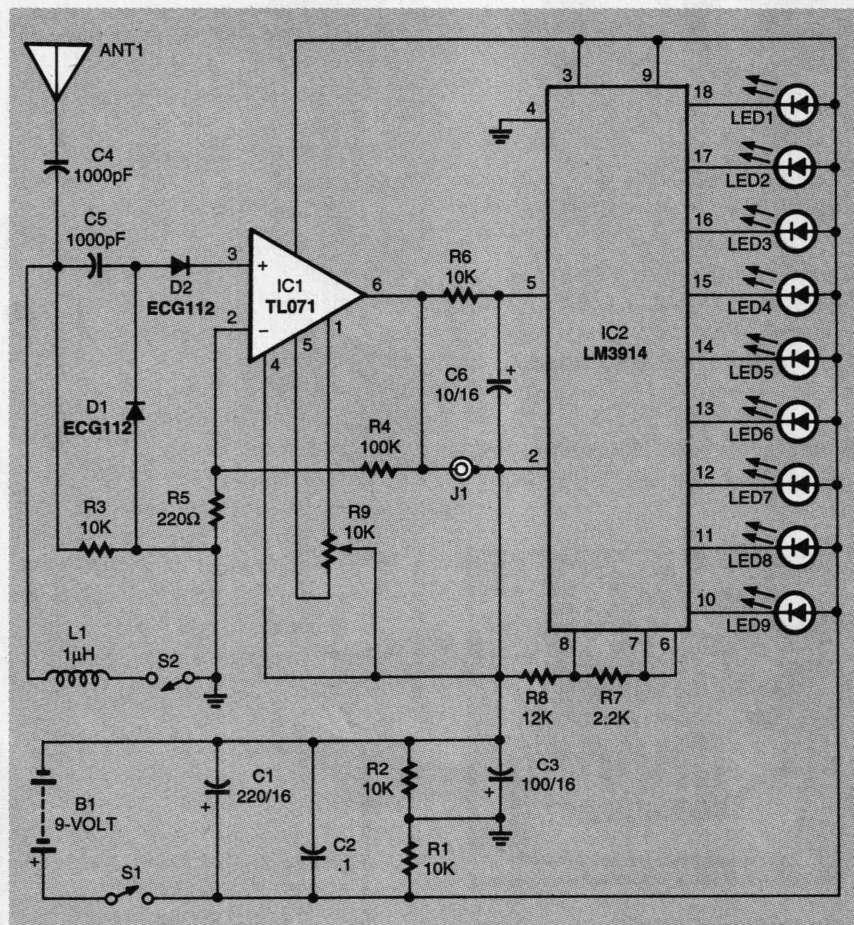
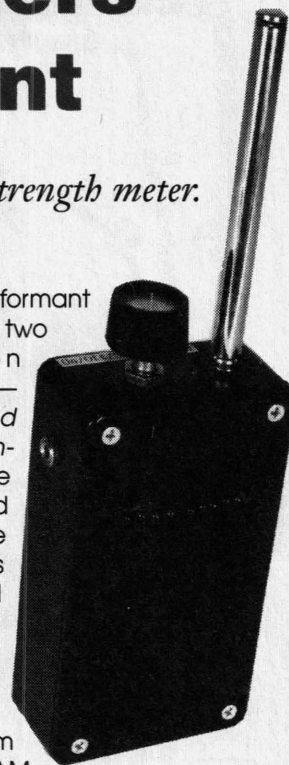


Fig. 1. The RF Informant is a wideband receiver that can show the relative strength of the received signal on a series of LEDs. An earphone can also be used to listen to the received signal. Closing S2 can enhance sensitivity in the FM band.

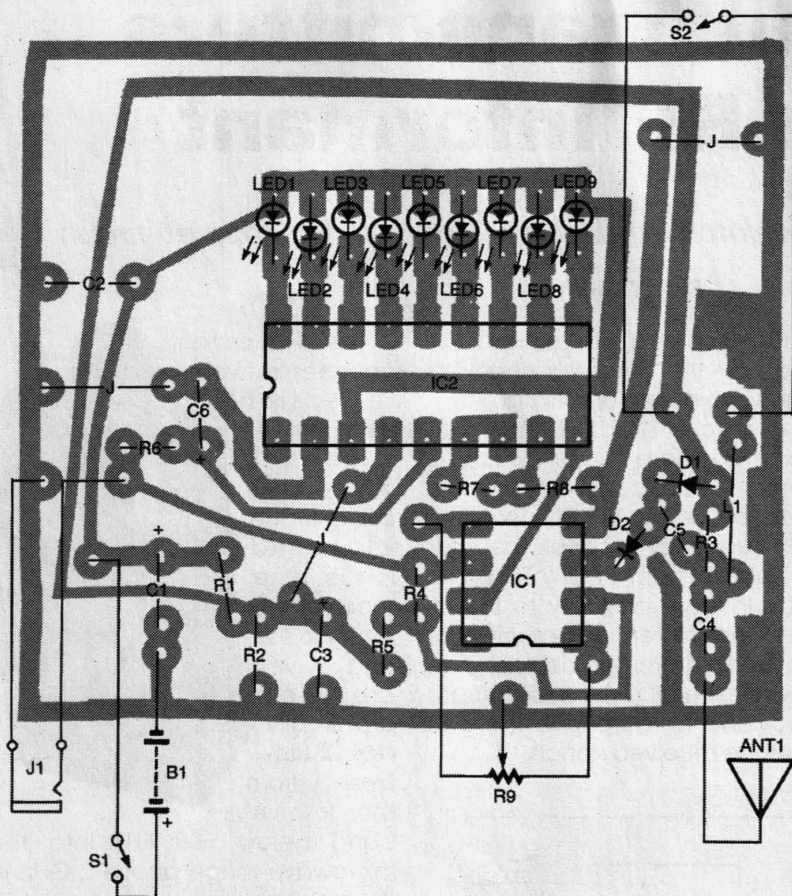


Fig. 2. The RF Informant must be built on a PC board; use this parts-placement diagram if you are using the foil pattern provided in this article. Don't forget to install the jumpers where indicated.

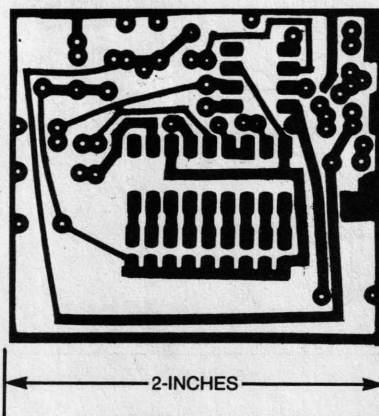
phone can be connected to J1 if you need to listen to the detected signal. Additionally, R6 and C6 smooth the signal.

The smoothed signal is then applied to the input of IC2, an LM3914 dot/bar display. That chip contains a resistor network and a set of comparators. Depending on the input voltage applied to pin 5, one or more LEDs will be turned on to display the relative voltage level. In the RF Informant, a bar display is selected by tying pin 3 to the positive supply voltage; at the lowest voltage, only LED9 will be illuminated. As the voltage increases, each LED comes on in turn until, at the highest voltage level, all nine devices are glowing. Resistors R7 and R8 set the reference voltage for a full-scale reading. Note that there are no current-limiting resistors for the LEDs; R7 and R8 limit the LED current as well.

Since IC1 requires a split power supply, R1, R2, and C3 create a

"ground" reference. Even though power is supplied by B1, a 9-volt battery, C1 and C2 filter any noise that might stray into the supply lines.

Building the RF Informant. Since the RF Informant uses high frequen-



Here's the foil pattern for the RF Informant. Using a single-sided board makes the board easy to etch and the project easy to build.

PARTS LIST FOR THE RF INFORMANT

SEMICONDUCTORS

IC1—TL071 operational amplifier, integrated circuit
 IC2—LM3914 Dot/bar display driver, integrated circuit
 DI, D2—ECG112 or similar silicon diode
 LED1-LED9—Light-emitting diode, red subminiature

RESISTORS

(All resistors are 1/4-watt, 5% units unless otherwise noted.)
 R1-R3, R6—10,000-ohm
 R4—100,000-ohm
 R5—220-ohm
 R7—2200-ohm
 R8—12,000-ohm
 R9—10,000-ohm potentiometer, panel mount (see text)

CAPACITORS

C1—220- μ F, 16-WVDC, electrolytic
 C2—0.1- μ F, ceramic-disc
 C3—100- μ F, 16-WVDC, electrolytic
 C4, C5—1000-pF, ceramic-disc
 C6—10- μ F, 16-WVDC, electrolytic

ADDITIONAL PARTS AND MATERIALS

L1—1- μ H inductor
 S1—Single-pole, single-throw switch (see text)
 S2—Single-pole, single-throw switch
 J1—Subminiature phone jack
 B1—9-volt battery
 ANT1—Telescoping antenna

Note: The following items are available from Quantum Research, 17919 77th Ave., Edmonton, AB, CA T5T 2S1: Etched and drilled PC board, \$10.00; Kit of all parts, enclosure, and PC board, \$69.95; Assembled RF Informant, \$99.95. Please add \$5.00 for shipping. All prices are in US dollars. Canadian residents must add appropriate PST and GST.

cies, a printed-circuit board must be used. A foil pattern has been included for making your own board. If you don't want to etch a board, one is available from the source given in the Parts List.

Follow the parts-placement diagram shown in Fig. 2 for the locations of the various components. While there is no necessary order when installing the parts, it is best to always

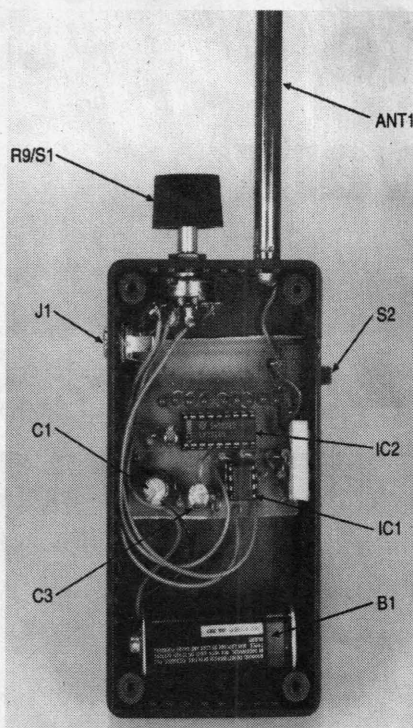


Fig. 3. The completed RF Informant fits in a case that's comfortable to hold. Note that in the author's prototype shown here, R9 and S1 are combined into a single unit. That way, a single control can be used for both switching the RF Informant on and off as well as adjusting the gain of IC1—a method used on almost all radios.

start with the smallest ones first; heat- and static-sensitive devices, such as semiconductors, should be saved for last. In light of that, start with the jumper wires. You can clip a short piece of lead from a resistor for use as a jumper. If you want to use sockets for the ICs, they should be mounted at this time, also.

The capacitors, resistors, and L1 are all mounted vertically. Note that C1, C3, and C6 are polarized components; double-check their orientation before installing them.

We are now ready for the semiconductors. Watch the orientation with those; be sure that they are facing the correct way before soldering them to the board. After mounting D1 and D2, install LED1–LED9. The light-emitting diodes must be mounted high enough off of the board so that they can be seen through the case cover. Most diodes have leads that will let you mount them so that the tips of the components are about 1 inch above the board. Keep them as long as possible.

Prepare the enclosure by drilling

holes for S1, S2, R9, J1, and ANT1. An example arrangement is shown in Fig. 3. Note that in that example (the author's prototype), R9 and S1 are one physical component. Using a switched potentiometer like that gives the RF Informant the operational feel of a standard radio. Install those components in the enclosure. If you absolutely can't locate a switched potentiometer, separate components could be used.

A window is needed on the cover for the LEDs. An easy way to "mill" a slot is to mark the location for the "window" and drill a row of holes across the cover. Once the holes are drilled, file the rough edges so that a slot is created. With the PC board temporarily placed in the case and the cover on, the LEDs should be visible through the slot.

Insulated wires are used to connect the rest of the components to the board. When everything is connected, the board is fixed in place with double-sided foam tape or a couple of dabs of silicone adhesive.

Finally, install the ICs in their sockets, making sure that they are plugged in the right way around. Snap a fresh 9-volt battery in place, and the RF Informant is ready for testing.

Testing and Use. Plug an earphone into J1 and turn on S1. Select S2 for wideband reception. With the antenna extended, you should hear some sounds, and one or more LEDs will probably be lit. Rotate R9 and note the behavior of the display; you will note that you can control whether the LEDs are on or off. With LED1 just at the point of turning on, the receiver is adjusted for maximum sensitivity.

To test the unit further you will need an RF transmitter such as a cellular phone, cordless phone, baby-room monitor, FM wireless mike, walkie-talkie, or any other similar device. Hold the RF Informant away from yourself with the antenna vertical. Sweep the unit in an arc. You will see that there are two directions that have the strongest reading. If the signal gains strength as you move in the direction of the signal, you should be getting closer to the signal source. If the signal

weakens, reverse your direction.

Adjust R9 for a reading on the LEDs that is not "pegged" to the limit. Continue to head in the direction of the strongest signal. You might need to correct your heading by sweeping the RF Informant in an arc to get a new bearing. You should also monitor the signal with an earphone plugged into J1. You might need to adjust the antenna length to pinpoint the RF source if it is quite strong.

If you are getting too much interference on the wideband setting, selecting the highband setting with S2 will usually give better reception—especially from an FM transmitter. Note that sources of interference can include light dimmers and fluorescent lighting.

Experience is the key to effectively using the RF Informant. Practice with hidden transmitters so that you can become familiar with the controls and monitoring techniques. You'll be amazed at how many sources of RF there are around you! Ω

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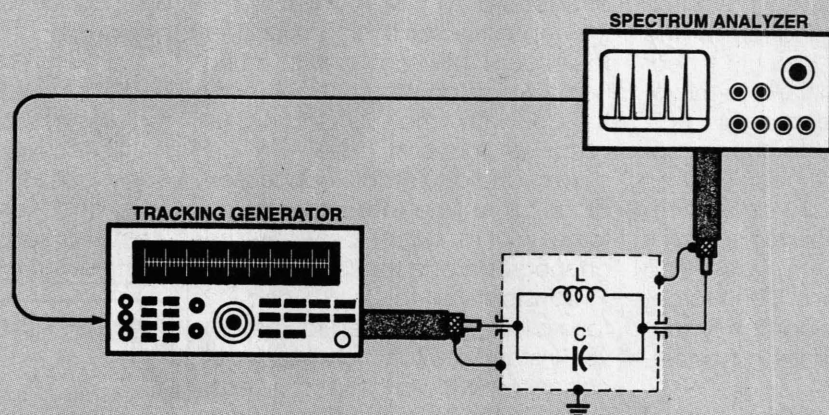
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July 1999, Electronics Now

Measuring Inductors and Capacitors at RF Frequencies



At low frequencies, measuring inductance and/or capacitance is no big deal, but get to RF frequencies and things get a lot more interesting.

JOSEPH J. CARR

The techniques needed to measure the values of inductors (L) and capacitors (C) at radio frequencies differs somewhat from what is needed to make the same measurements at low frequencies. In short, although similarities exist, the RF measurement is a bit more complicated. One of the reasons for that is that stray or "distributed" inductance and capacitance values of the test set-up will affect the results. Another reason is that capacitors and inductors are not ideal components, but rather all capacitors have some inductance, and all inductors have capacitance. In this article we will take a look at several effective methods for measuring inductance and capacitance at RF frequencies.

VSWR Method. When a load impedance is connected across an RF source, the maximum power transfer occurs when the load (Z_L) and source (Z_S) impedances are equal ($Z_L = Z_S$). If those impedances are not equal, then the *voltage standing wave ratio* (VSWR) will indicate the degree of mismatch. We can use that phenomenon to measure values of inductance and capacitance using the scheme

shown in Fig. 1A. The instrumentation required includes a signal generator or other signal source, and a VSWR meter or VSWR analyzer.

Some VSWR instruments require a transmitter for excitation, but others will accept the lower signal levels that can be produced by a signal generator. An alternative device is an SWR-analyzer type of instrument. It contains the signal generator and VSWR meter, along with a frequency counter to be sure of the actual test frequency. Whatever signal source is used, however, it must have a variable output frequency. Further, the frequency readout must be accurate (the accuracy of the method depends on knowing the actual frequency).

The load impedance inside the shielded enclosure consists of a non-inductive resistor (R) that has a resistance equal to the desired system impedance resistive component (50 ohms in most RF applications, and 75 ohms in television and video). An inductive reactance (X_L) and a capacitive reactance (X_C) are connected in series with the load. The circuit containing a resistor, capacitor, and inductor simu-

lates an antenna-feedpoint impedance. The overall impedance is:

$$Z_L = \sqrt{R^2 + (X_L - X_C)^2} \quad (1)$$

Note the reactive portion of Equation 1. When the condition $|X_L| = |X_C|$ exists, the series network is at resonance, and VSWR is minimum (see Fig. 1B). This gives us a means for measuring the values of the capacitor or inductor, provided that the other is known. That is, if you want to measure a capacitance, then use an inductor of known value. Alternatively, if you want to know the value of an unknown inductor, use a capacitor of known value.

Using the test set-up in Fig. 1A, adjust the frequency of the signal source to produce minimum VSWR.

1. For finding an inductance from a known capacitance:

$$L_{\mu H} = \frac{10^{12}}{4\pi^2 f^2 C_{PF}} \quad (2)$$

Where:

$L_{\mu H}$ = inductance in microhenrys (μH)
 C_{PF} is the capacitance in picofarads (pF)

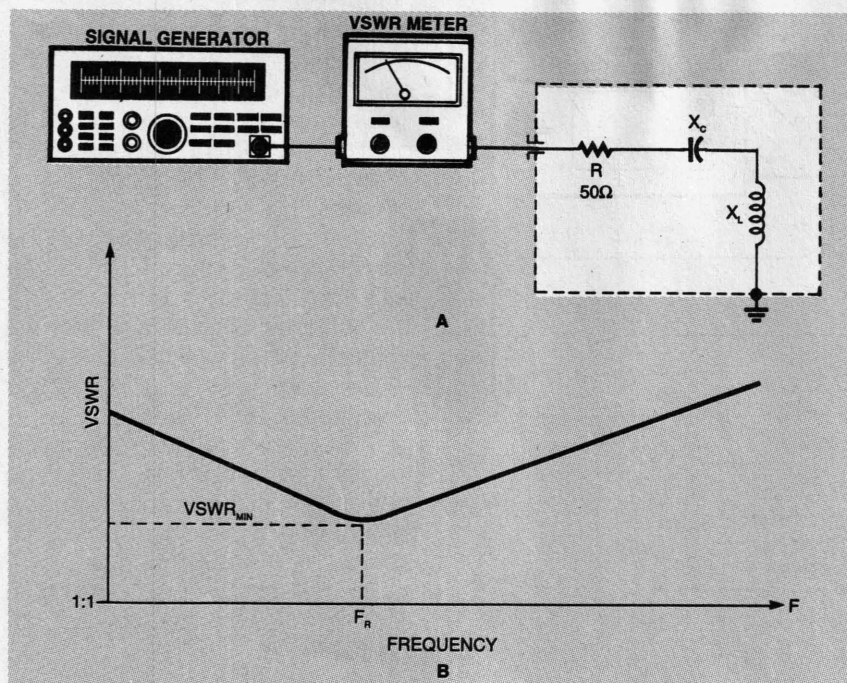


Fig. 1. The VSWR method for measuring L and C is shown in A while the VSWR-vs.-frequency curve is shown in B.

f is the frequency in hertz (Hz)

2. For finding a capacitance from a known inductance:

$$C_{pF} = \frac{10^{12}}{4\pi^2 f^2 L_{\mu H}} \quad (3)$$

The accuracy of this approach depends on how accurately the frequency and the reactance are known, and how accurately the minimum VSWR frequency can be found.

Voltage Divider Method. A resistive voltage divider is shown in Fig. 2A. This circuit consists of two resistors (R_1 and R_2) in series across a voltage source V . The voltage drops across R_1 and R_2 are V_1 and V_2 , respectively. We know that either voltage drop is found from:

$$V_X = \frac{V R_X}{R_1 + R_2} \quad (4)$$

Where: V_X is V_1 and R_X is R_1 or, V_X is V_2 and R_X is R_2 , depending on which voltage drop is being measured.

We can use the voltage divider concept to find either inductance or capacitance by replacing R_2 with the unknown reactance.

Consider first the inductive case. In Fig. 2B resistor R_2 has been replaced by an inductor (L). Resistor R_1 is the inductor series resistance. If we measure the voltage drop across R_1 (i.e. "E" in Fig. 2B), then we can calculate the inductance from:

$$L = \frac{R}{2\pi f} \times \sqrt{\left(\frac{V}{E}\right)^2 - \left(1 + \frac{R_s}{R_1}\right)^2} \quad (5)$$

As can be noted in Equation 5, if $R_1 \gg R_s$, then the quotient R_s/R_1 becomes negligible.

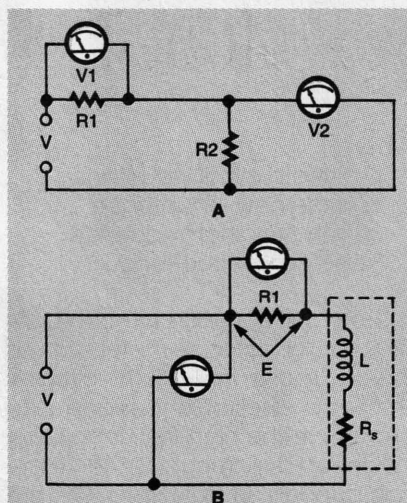


Fig. 2. A simple voltage divider is shown in A and a reactance voltage divider is in B.

In capacitors the series resistance is typically too small to be of consequence. We can replace L in the model of Fig. 2B with a capacitor, and again measure voltage E . The value of the capacitor will be:

$$C = \frac{2\pi f \times 10^6}{R \times \sqrt{\left(\frac{V}{E}\right)^2 - 1}} \quad (6)$$

The value of resistance selected for R_1 should be approximately the same order of magnitude as the expected reactance of the capacitor or inductor being measured. For example, if you expect the reactance to be, say, between 1000- and 10,000-ohms at some frequency, then select a resistance for R_1 in this same range. That will keep the voltage values manageable.

Signal Generator Method. If the frequency of a signal generator is accurately known, then we can use a known inductance to find an unknown capacitance, or a known capacitance to find an unknown inductance. Figure 3 shows the test set-up for this option. The known and unknown components (L and C) are connected together inside a shielded enclosure. The parallel-tuned circuit is lightly coupled to the signal source and the display through very low-value capacitors (C_1 and C_2). The rule is that the reactance of C_1 and C_2 should be very high compared with the reactances of L and C at resonance.

The signal generator is equipped with a 6-dB resistive attenuator in order to keep its output impedance stable. The output indicators should be any instrument that will read the RF voltage at the frequency of resonance. For example, you could use either an RF voltmeter or an oscilloscope.

The procedure requires tuning the frequency of the signal source to provide a peak output-voltage reading on the voltmeter or scope. If the value of one of the components (L or C) is known, then the value of the other can be calculated using Equation 2 or 3, as appropriate.

Alternate forms of coupling are 41

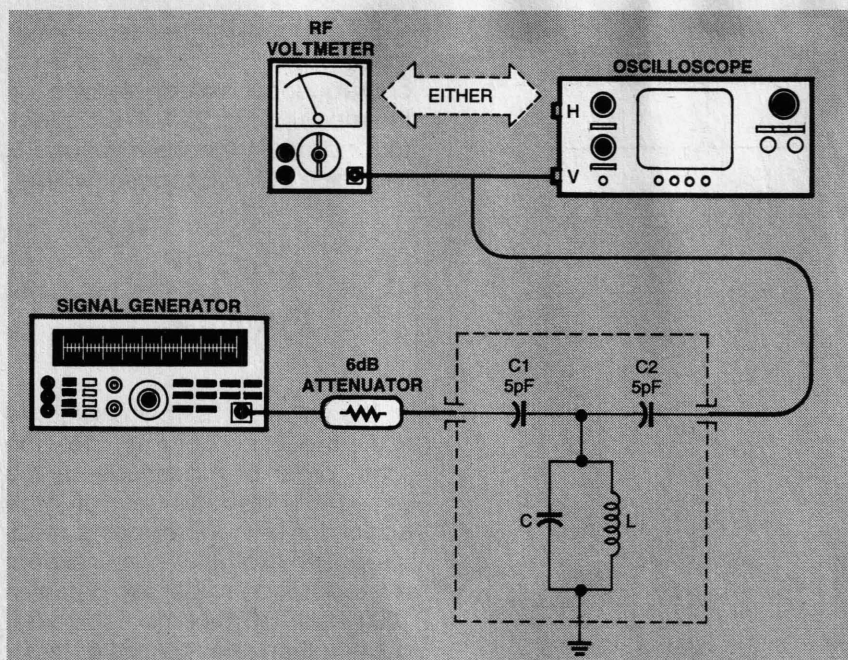


Fig. 3. If the frequency output of a signal generator is accurately known, we can use a known inductance to find an unknown capacitance, or vice versa.

shown in Fig. 4. In either case, the idea is to isolate the instruments from the L and C elements. In Fig. 4A the isolation is provided by a pair of high-value (10,000-ohm to 1 megohm) resistors, R1 and R2. In Fig. 4B the coupling and isolation is pro-

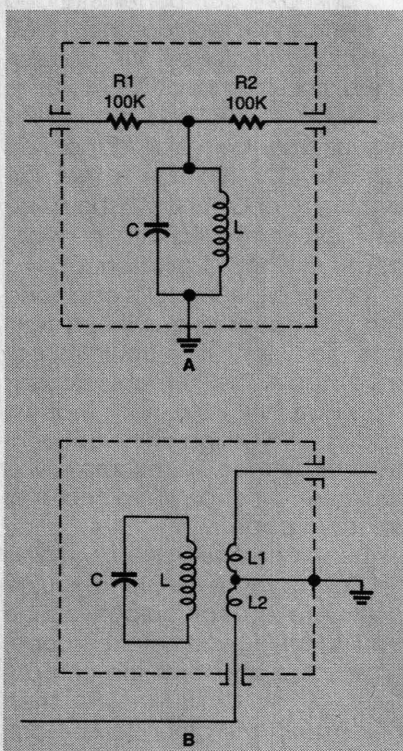


Fig. 4. Different ways to couple L and C elements to the test instruments.

vided by a one- or two-turn link winding over the inductor. The links and the main inductor are lightly coupled to each other.

Frequency-Shifted Oscillator Method.

The frequency of a variable-frequency oscillator (VFO) is set by the combined action of an inductor and a capacitor. We know that a change in either capacitance or inductance produces a frequency change equal to the square of the component ratio. For example, for an inductance change:

$$L2 = L1 \times \left[\left(\frac{F1}{F2} \right)^2 - 1 \right] \quad (7)$$

Where:

- L1 is the original inductance
- L2 is the new inductance
- F1 is the original frequency
- F2 is the new frequency

From this equation we can construct an inductance meter such as the one shown in Fig. 5. This circuit is a Clapp oscillator designed to oscillate in the high-frequency (HF) range up to about 12 MHz. The components L1, C1, and C2 are selected to resonate at some frequency. Inductor L1 should be of

the same order of magnitude as L_x . The idea is to connect the unknown inductor across the test fixture terminals. Switch S1 is set to position "b" and the frequency (F1) is measured on a digital frequency counter. The switch is then set to position "a" in order to put the unknown inductance (L_x) in series with the known inductance (L1). The oscillator output frequency will shift to F2. When we know L1, F1 and F2 we can apply Equation (7) to calculate L_x (L2 in Equation 7).

If we need to find a capacitance, then modify the circuit to permit a capacitance to be switched into the circuit across C1 instead of an inductance as shown in Fig. 5. Replace the "L" terms in Equation (7) with the corresponding "C" terms.

Using RF Bridges. Most RF bridges are based on the DC Wheatstone bridge circuit (see Fig. 6). In use since 1843, the Wheatstone bridge has formed the basis for many different measurement instruments. The *null condition* of the Wheatstone bridge exists when the voltage drop of R1/R2 is equal to the voltage drop of R3/R4. When the condition $R1/R2 = R3/R4$ is true, then the voltmeter (M1) will read zero. The basic measurement scheme is to know the values of three of the resistors, and use them to measure the value of the fourth. For example, one common scheme is to connect the unknown resistor in place of R4, make R1 and R3 fixed resistors of known value, and R2 is a calibrated potentiometer marked in ohms. By adjusting R2 for the null condition, and then reading its value, we can use the ratio $(R2 \times R3)/R1 = R4$.

The Wheatstone bridge works well for finding unknown resistances from DC to some relatively low RF frequencies, but to measure L and C values at higher frequencies we need to modify the bridge. Three basic versions are used: Maxwell bridge (Fig. 7), Hay bridge (Fig. 8), and Schering bridge (Fig. 9).

Maxwell Bridge. The Maxwell bridge is shown in Fig. 7. The null condition for this bridge occurs when:

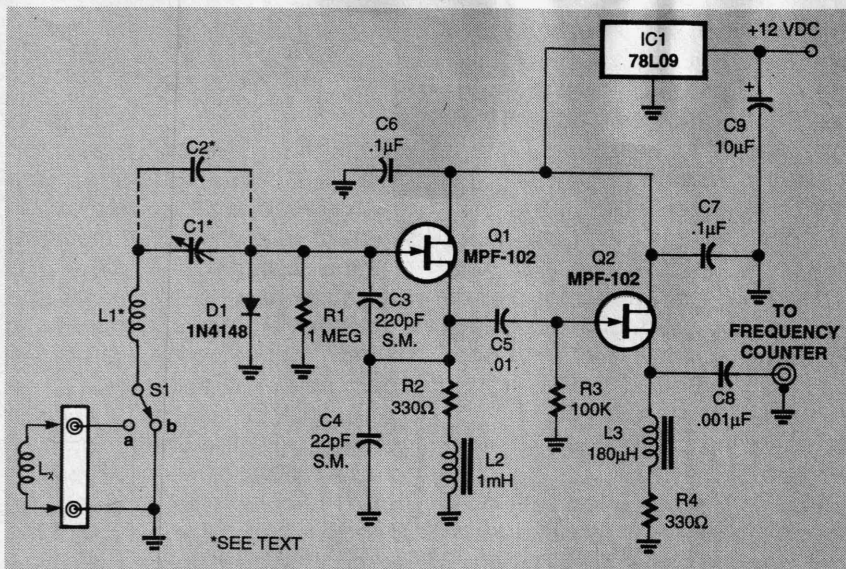


Fig. 5. This circuit uses the frequency-shift method to measure the value of an unknown inductance. It is easily modified to measure an unknown capacitor instead.

$$L1 = R2 \times R3 \times C1 \quad (8)$$

and

$$R4 = \frac{R2 \times R3}{R1} \quad (9)$$

The Maxwell bridge is often used to measure unknown values of inductance (e.g. $L1$) because the balance equations are totally independent of frequency. The bridge is also not too sensitive to resistive losses in the inductor (a failing of some other methods). Additionally, it is much easier to obtain calibrated standard capacitors for $C1$ than it is to obtain standard inductors for $L1$. As a result, one of the principal

uses of this bridge is inductance measurements.

Maxwell bridge circuits are often

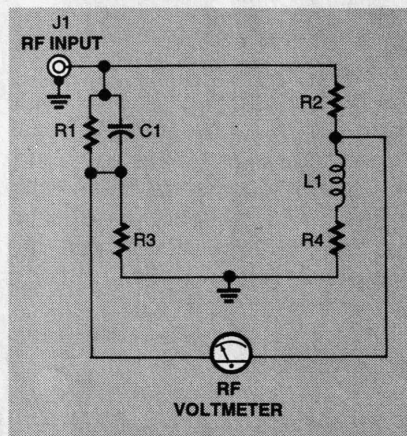


Fig. 7. In the Maxwell bridge, the balance equations are independent of the frequency.

used in measurement instruments called *Q-meters*, which measure the quality factor (Q) of inductors. The equation for Q is, however, frequency sensitive:

$$Q = 2 \times \pi \times F \times R1 \times C1 \quad (10)$$

Where: F is in Hertz, $R1$ in ohms, and $C1$ in farads.

Hay Bridge. The Hay bridge (see Fig. 8) is physically similar to the Maxwell bridge, except that the $R1/C1$ combination is connected in series rather than parallel. Unlike the Maxwell bridge, the Hay bridge is

frequency sensitive. The balance equations for the null condition are also a little more complex:

$$L1 = \frac{R2 \times R3 \times C1}{1 + \left[\frac{1}{Q}\right]^2} \quad (11)$$

$$R4 = \left[\frac{R2 \times R3}{R1}\right] \times \left[\frac{1}{Q^2 + 1}\right] \quad (12)$$

Where:

$$Q = \frac{1}{\omega \times R1 \times C1} \quad (13)$$

The Hay bridge is used for measuring inductances with high Q figures, while the Maxwell bridge is best with inductors that have a low Q value.

Note: A frequency-independent version of Equation (11) is possible when Q is very large (i.e. >100):

$$L1 = R2 \times R3 \times C1 \quad (14)$$

Schering Bridge. The Schering bridge circuit is shown in Fig. 9. The balance equations for the null condition are:

$$C3 = \frac{C2 \times R1}{R2} \quad (15)$$

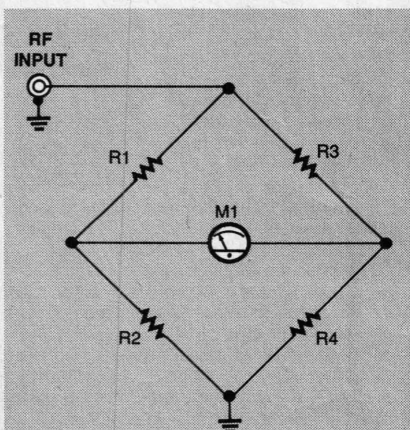


Fig. 6. The Wheatstone bridge has been used since 1843.

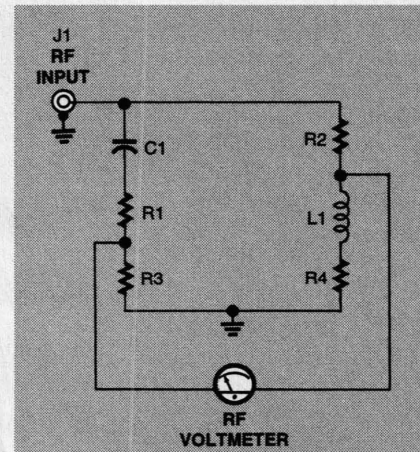


Fig. 8. The Hay bridge is used to measure inductances with high Q values..

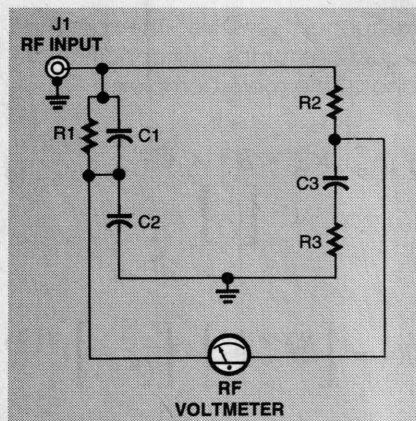


Fig. 9. The Schering bridge is used to find the capacitance and power factor of capacitors.

and

$$R3 = \frac{C2 \times R1}{R2} \quad (16)$$

The Schering bridge is used primarily for finding the capacitance and the power factor of capacitors. In the latter applications no actual R3 is connected into the circuit, making the series resistance of the capacitor being tested (e.g. C3) the only resistance in that arm of the bridge. The capacitor's Q factor is found from:

$$Q_{C3} = \frac{1}{\omega \times R1 \times C1} \quad (17)$$

Finding Parasitic Capacitances and Inductances. Capacitors and inductors are not ideal components. A capacitor will have a certain amount of series inductance (called "parasitic inductance"). This inductance is created by the conductors in the capacitor, especially the leads. In older forms of capacitor, such as the wax-paper dielectric devices used prior to about 1960, the series inductance was very large. Because the inductance is in series with the capacitance, it forms a series-resonant circuit.

Figure 10 shows a test set-up for finding the series-resonant frequency. A tracking generator is a special form of sweep generator that is synchronized to the frequency sweep of a spectrum analyzer. They are used with spectrum analyzers in order to perform stimulus-response measurements such as Fig. 10.

The nature of a series-resonant circuit is to present a low impedance at the resonant frequency, and a high impedance at all frequencies removed from resonance. In this case (Fig. 10), that impedance is across the signal line. The display on the spectrum analyzer will show a pronounced, sharp dip at the frequency where the capacitance and the parasitic inductance are resonant.

The value of the parasitic series inductance is:

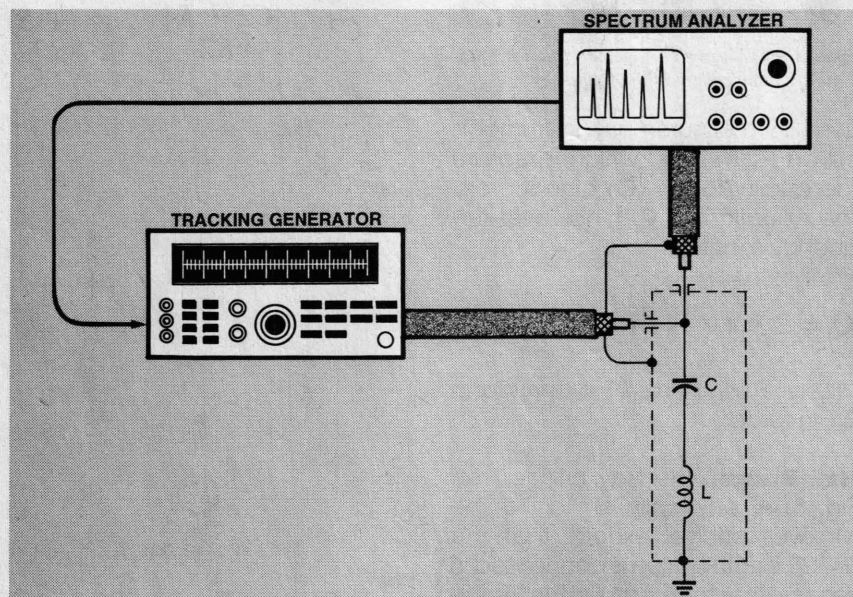
$$L = \frac{1}{2^2 \pi^2 f^2 C} \quad (18)$$

Inductors are also less than ideal. The adjacent turns of wire form small capacitors, which when summed up can result in a relatively large capacitance value. The illustration that appears at the beginning of this article shows a method for measuring the parallel capacitance of an inductor.

Because the capacitance is in parallel with the inductance, it forms a parallel-resonant circuit. Those circuits will produce an impedance that is very high at the resonant frequency, and very low at frequencies removed from resonance. In the lead illustration, the inductor and its parasitic parallel capacitance are in series with the signal line, so will (like the other circuit) produce a pronounced dip in the signal at the resonant frequency. The value of the parasitic inductance is:

$$C = \frac{1}{2^2 \pi^2 f^2 L} \quad (19)$$

Conclusion. There are other forms of bridges, and other methods, for measuring L and C elements in RF circuits, but those discussed here are the most practical. That's especially true if you do not own or have access to specialized instruments. Ω



THE COLLECTED WORKS OF MOHAMMED ULLYES FIPS

#166—By Hugo Gernsback. Here is a collection of 21 April Fools Articles, reprinted from the pages of the magazines they appeared in, as a 74-page, 8 1/2 x 11-inch book. The stories were written between 1933 and 1964. Some of the devices actually exist today.

Others are just around the corner. All are fun and almost possible. Stories include the Cordless Radio Iron, The Visi-Talkie, Electronic Razor, 30-Day LP Record, Teleyeglasses and even Electronic Brain Servicing. Get your copy today. Ask for book #166 and include \$9.99 (includes shipping and handling) in the US (First Class), Canada and Overseas (surface mail), and order from CLAGGK Inc., P.O. Box 4099, Farmingdale, NY 11735-0793. Payment in US funds by US bank check or International Money Order. Allow 6-8 weeks for delivery.

MA05

Q & A

READERS' QUESTIONS, EDITORS' ANSWERS
CONDUCTED BY MICHAEL A. COVINGTON, N4TMI

ER...That Was A Joke!

Our April Fools' Day joke, "The EC909-12 Analog Microprocessor," apparently fooled a number of people too well—we've received a flood of mail about it.

Note that Ecraf is *farce* spelled backward, no address is given for any person or company involved, the claims are outlandish (such as emission of huge amounts of light—enough to light a city—at tiny currents), and the last paragraph of the article says it's due for release on April 1, our national day of tomfoolery.

Two items in "Prototype" (the DVD rewinder and the WIG-WOM) were also put-ons and also specifically mentioned April 1. But the equally improbable-sounding quantum tunneling transistor (pp. 42-43, 50) is real; see www.sandia.gov/media/quantran.htm for more information.

April Fool's jokes are a decades-old tradition at **Electronics Now**; our founder, Hugo Gernsback, published facetious articles under the name of Mohammed Ulysses Fips. Interestingly, though improbable at the time, some of the concepts in those articles—such as an optical audio disc—were uncannily predictive of the future. And that's the problem—one person's fiction can easily be another person's invention.

Ultrasonic Listener

Q Where can I find plans or a kit for a circuit to allow me to hear ultrasonic sounds such as bats' squeaks?—L. S., Newton, MA

A I'm not aware of a complete, published project or kit, but Fig. 1 shows a circuit you can experiment with. I used it successfully to listen to the squeaks of a domesticated rat.

The principle involved is *heterodyning* (mixing), a process that gives you the sum and difference of the original frequencies. For example, if you mix a 25-

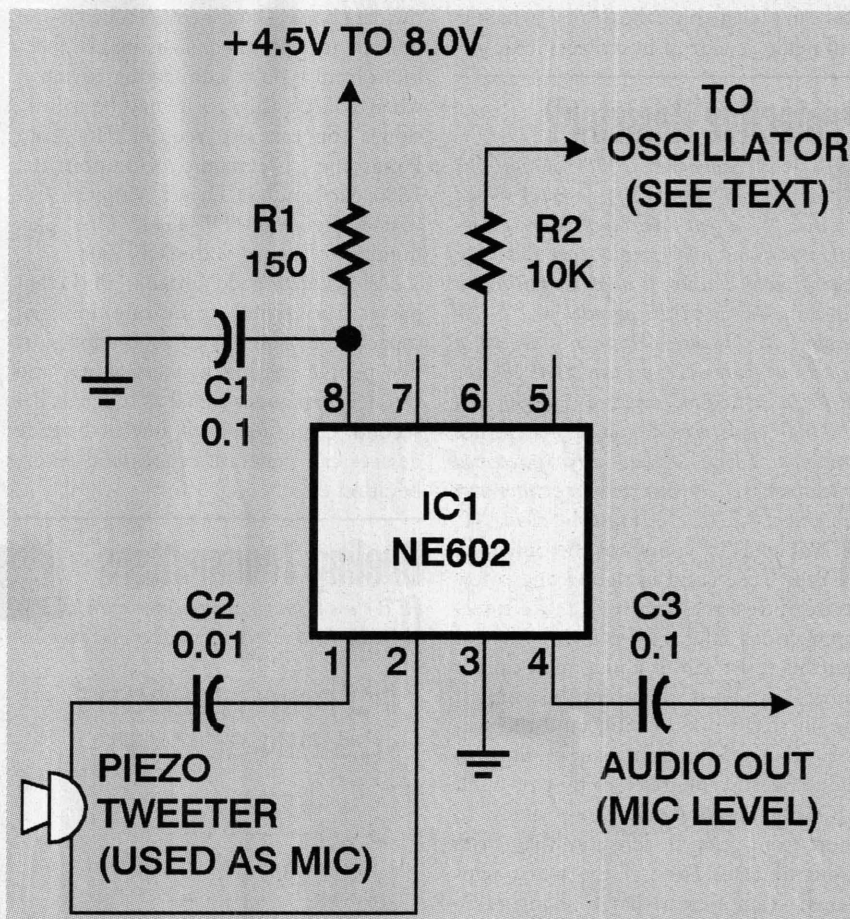


FIG. 1—HERE'S A GOOD STARTING POINT for an ultrasonic-listening circuit. The NE602 mixes 25-kHz ultrasound with a 20-kHz oscillator signal, yielding 5-kHz audio. If you can't find the now-discontinued NE602, the SA602 (still being made) is a direct replacement.

kHz incoming signal with the output of a 20-kHz oscillator, you get additional signals at $25 + 20 = 45$ kHz and $25 - 20 = 5$ kHz; the latter is audible.

Heterodyning takes place when one signal is multiplied by the other; that is, when the amplification of one signal is proportional to the signal level of the other. Any nonlinear amplifier will do this, but the NE602 chip shown in the circuit uses a sophisticated mixer called a *Gilbert cell* whose advantage is that not much of the original signal appears at the output.

The NE602 was designed for radio

receiver front-ends, and it works with low-level signals, so the microphone connects directly to its input. As shown, the microphone is connected differentially across pins 1 and 2. If one side of the microphone is grounded, you can connect the ungrounded microphone lead to C2 and add a capacitor, equal to C2, from pin 2 to circuit ground. Although the NE602 provides some amplification, the output signal is still microphone-level.

The design of the oscillator is up to you; its frequency should be adjustable so you can tune it near the ultrasonic

signal you want to hear. From R2 it goes directly into the base of a transistor. Logic-level (5-volt) oscillators work well; I used a signal generator for experiments, but a 555 chip would be a good choice when building a self-contained unit.

Note that the NE602 is no longer being manufactured, but should still be available from a number of sources. If you can't locate one, the SA602, which is still made, is a drop-in replacement.

Protective Resistor?

Q In your September 1998 column, the answer and Fig. 1 (Fig. 2 here) should specify a "build-out" resistance on any external connections to PC printer port pins other than ground. Failure to do so is inviting an expensive repair to the parallel port if, for instance in this case, Q1 were to develop a defect and apply +12 volts to pin D0. For the IRF510, a resistor between 10,000 and 100,000 ohms would provide the needed protection. I hope you have already received at least a dozen comments about printer port protection.—J. C. H., Winston-Salem, NC

A No, actually, yours was the only one. Your basic point is a good one; a few strategic resistors can help make interconnections safer. However, whether to put this resistor in is a judgment call; in general, we want to present the simplest circuit that works reliably, and the scenario that you describe is quite uncommon. Because of the way they're built, MOSFETs are much more likely to short from drain to substrate than from drain to gate. Even if the worst happened, it's not clear that applying +12V would actually damage a modern parallel port, which includes some protective resistance of its own; higher voltages certainly would, of course.

Because this circuit drives a relay, signal speed is not an issue, but in general, adding relays will slow down the charging of cable capacitance and gate capacitance. This can be good or bad. Small series resistors (on the order of 330 ohms) often make a long data cable work more reliably by reducing crosstalk. A resistor as large as 10,000 ohms, though, should be placed close to Q1, not at the PC, in order not to weaken the signal going into the cable.

Bucket-Brigade Devices

Q Back in the 1970s, there were audio delay chips called bucket-brigade analog delays.

Are they still being made? Are books about them available?—C. K. S., Waymart, PA

A Bucket-brigade delay devices are still made by Panasonic (Matsushita) and are available from Panasonic distributors, one of whom is Digi-Key, 701 Brooks Ave. S., Thief River Falls, MN 56701; Tel: 800-344-4539; Web: www.digikey.com. Digi-Key also sells the data book. Data sheets can be viewed online at www.mec.panasonic.co.jp/e-index.html (you're connecting to Japan when you do this, so it may be a little slow). You can also request data from Panasonic Electronic Components, 1600 McCandless Drive, Milpitas, CA 95035; Tel: 408-946-4311. One part number to look at is the MN3008.

A bucket-brigade device is an IC that passes the signal from one internal capacitor to another in succession, just like people passing buckets along, and can delay an audio signal as much as 0.4 second. Using feedback, bucket-brigade devices can generate echoes and reverberation effects.

Dueling Transmitters

Q We use a simple transmitter with an FM radio and outdoor speaker to monitor

activity on the internal phone line of the coal mine where I work. Teenagers in the area have started a community FM-broadcast station which interferes with our reception no matter where we tune the transmitter and radio. Can our transmitter be modified to tune somewhere in the 30-50 MHz range so we can use a scanner to monitor the activity in the mine?—D. B., Jesse, WV

A Probably, but let's solve the *real* problem instead. What you're experiencing is a conflict between two unlicensed transmitters, yours and theirs. Unlicensed transmitters are not allowed to interfere with licensed ones, so if the teenagers are keeping you from receiving FM broadcast stations (not just your own transmitter), you have a clear case against them. Contact the teenagers, explain the problem, and explain that you will report them to the FCC if they don't clean up their signal. Or, if you prefer, contact the Federal Communications Commission, Tel: 888-225-5322 and report the problem directly.

However, I believe that safety-critical communications should never rely on an unlicensed transmitter; you just have no assurance that your signals will not be interfered with. Consider relaying the

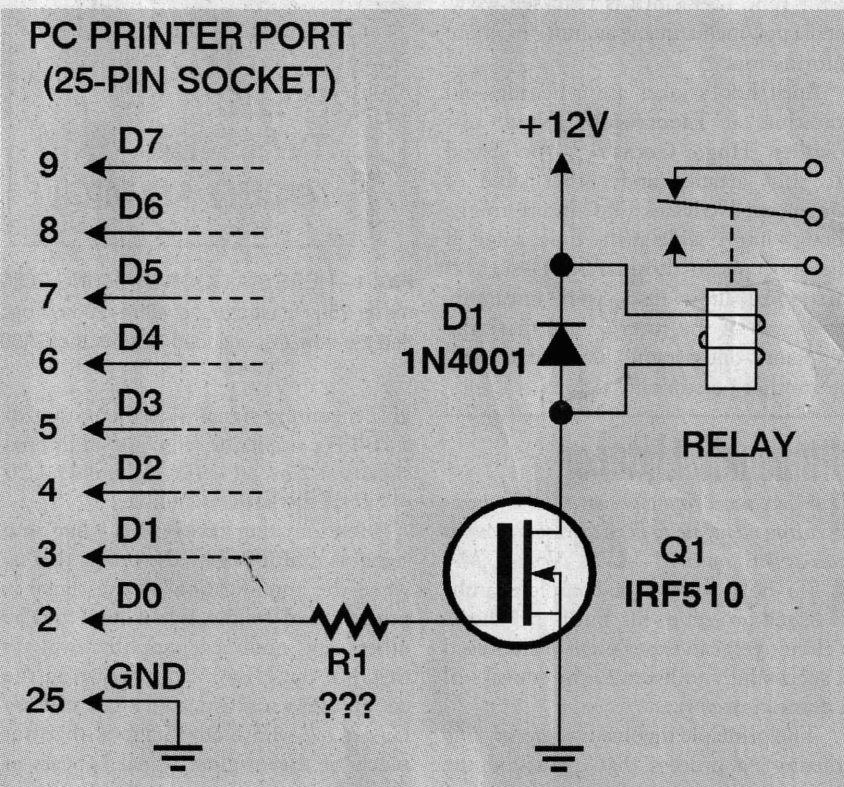
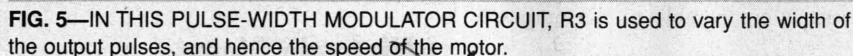


FIG. 2—ADDING A RESISTOR between the printer port and any external circuit connected to it could add protection in case of a failure, but is it really needed? See text.



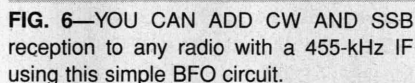
The BFO is a separate circuit. You can use the oscillator shown in Fig. 6; it originally appeared in **Popular Electronics**, May 1990. Note that it does not connect directly to the radio; instead, you place the coupling wire near or around the detector diode. Once you've tuned the BFO initially, you can probably fine-tune individual stations with the radio's tuning knob without making further adjustments to the BFO.

Q *I would like to convert the electronic clock-timer of an oven to operate on 50-Hz power. It uses an HD 614042 IC. — R. N. L., Curacao, Netherlands Antilles*

A We assume you've solved all the other voltage problems, so the clock is working correctly but running at % of normal speed. We're not familiar with the HD 614042 chip, but if you can find out where it receives its 60-Hz input, perhaps you can build the oscillator in Fig. 4 and supply the 60-Hz signal from that rather than from the line.

Q *I would like to install an electric drive on an adult tricycle for use in a retirement community, with a solid-state "chopper" speed control. I've been advised that such a controller might be constructed using the 555 timer in PWM mode. Can you help? — R. L., Kissimmee, FL*

A A motor-speed control circuit that you can probably adapt for your purposes is shown in Fig. 5. In the prototype, we used an IRF510 and a small



tape-recorder motor; for your big motor, you'll want to use a number of MOSFETs in parallel. The 555 operates as a pulse-width modulator (PWM), producing a squarewave that is positive for an adjustable percentage of the time. Thus, the average voltage at the motor can be varied without wasting any energy in a rheostat.

As always, we welcome your questions. Please write to Q&A, **Electronics Now Magazine**, 500 Bi-County Blvd., Farmingdale, NY 11735. The most interesting ones are answered in print. Please be sure to include plenty of background information (we'll shorten your letter for publication). *If you are asking about a circuit, please include a complete diagram.* Due to the volume of mail, we regret that we cannot give personal replies.

The text contains a goodly number of practical music projects most often requested by musicians. All the projects are relatively low-in-cost to build readily-available component. The project categories are music and MIDI.

Please allow 6-8 weeks for delivery.

THE DISCONE ANTENNA DESCRIBED here was designed to operate over a frequency range of 700 to 2000 MHz. It is a very small antenna, but very effective. It is vertically polarized, and has an omni-directional radiation pattern.

The antenna's principle elements are a flat conducting disc mounted horizontally atop but insulated from a conducting cone. The disc diameter is about 0.17 wavelength at the lowest desired operational frequency, and the cone has a length of 0.25 wavelength on the side. The discone's impedance is 50 ohms, so it can be fed by a 50-ohm coaxial cable. The outer (shield) conductor is connected to the cone, the center conductor is connected to the disc. The antenna's actual impedance varies depending on the cone's angle, frequency, and disc-to-cone spacing. Nevertheless, discone dimensions are not very critical for optimum performance.

Figure 1-a is an idealized sketch of a true discone antenna and its basic dimensions. It was determined that the disc diameter should be 3 inches and the length of a side element of the cone should be $4\frac{1}{4}$ inches. The angle θ (Fig. 1-b) could be any angle between 25 to 40 degrees, so 30 degrees was selected for a practical reason. The space created by the insulating washer (S) will be $\frac{1}{8}$ of the inside diameter of the brass tube (Fig. 2) or about $\frac{1}{8}$ inch (thickness is not critical).

Construction

A piece of $\frac{5}{8}$ -inch brass tube is used to support the discone and is the outer conductor of the feedline (Figs. 1-b and 2). This tube has an inside diameter of $\frac{19}{32}$ inch. With commonly available $\frac{1}{4}$ -inch brass rod used for an inner conductor, the section of coaxial line that results has a 52-ohm impedance.

The exact impedance is not too critical and less than 10% variation in impedance should not cause reception problems. The length of the brass tube is up to the discretion of the builder. The loss the added length in-

BUILD THIS DISCONE VHF-UHF ANTENNA



The discone is a popular wide-band antenna for VHF and UHF. This article shows you how to build a very efficient one for your UHF scanner.

WILLIAM SHEETS, K2MQJ and RUDOLF F GRAF, KA2CWL

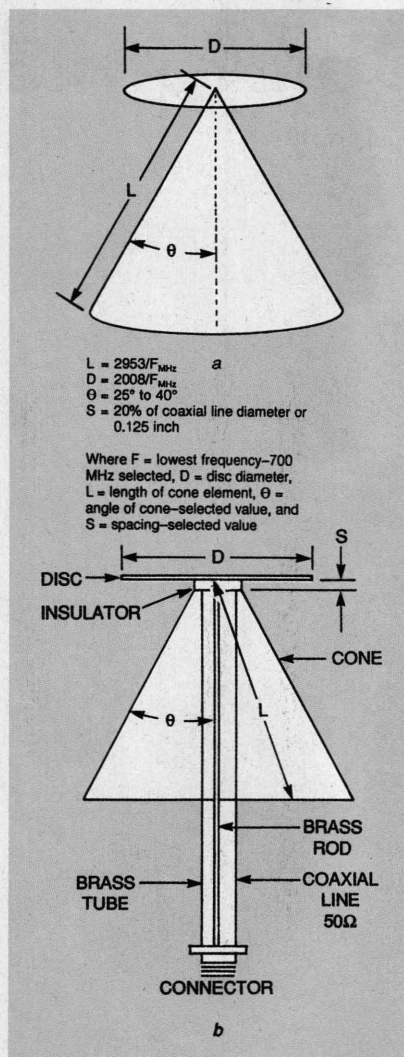


FIG. 1—DISCONE ANTENNA DIMENSIONS: a-view of idealized discone antenna and, b-view of assembled antenna specified in this article. Equations can be used for assembled antenna and provide excellent reception results.

introduces is negligible. A 5½-inch length of brass tube was used in the discone illustrated here, but up to about two feet of tubing should present no problems. Longer lengths will require some mechanical modifications in order to ensure that the line geometry remains concentric and reasonably rigid. This becomes a construction problem and should be avoided.

Theoretically the cone of the discone should come to a point. However, it can be truncated to allow the brass tube to be soldered to it. The disc is fastened to the brass rod (Fig. 3) by a screw which fits into a tapped

hole in the center conductor. A shoulder insulator made from plastic faucet washers keeps the brass rod concentric with the inner wall of the brass tube and provides a spacing between the disc and cone of about 0.125 inch. The bottom end of the line section is soldered to a type N UHF connector. A small clamp or U bolt can be used to mount the antenna to a mast.

Sheet metal work

The disc and cone were cut from .019 gauge copper flashing stock (Fig. 2) purchased from a local plumbing supply house. Since the angle selected is 30 degrees, a half-circle pattern is needed to form the cone. Cut the cone and disc according to pattern. Allow a little overlap tab

as shown in Fig. 2 to allow for soldering. Use shears and wear heavy gloves as copper tends to cut with sharp razor-like edges. File all edges smooth.

The cone is formed by first drawing radial lines on the inside surface, bending the pattern a little at each line around a block of wood or steel, and repeating the process until the pattern edges meet. The cone should be a fairly good, even, circular shape. Make sure the hole at the top will fit the 5/8-inch brass tube snugly. Clean the edges and soldering surfaces with fine (No. 0) steel wool. Clamp the edges of the cone together with the tab underneath and solder using 60/40 solid-core solder and a liquid flux. Next, clean the brass tube with fine

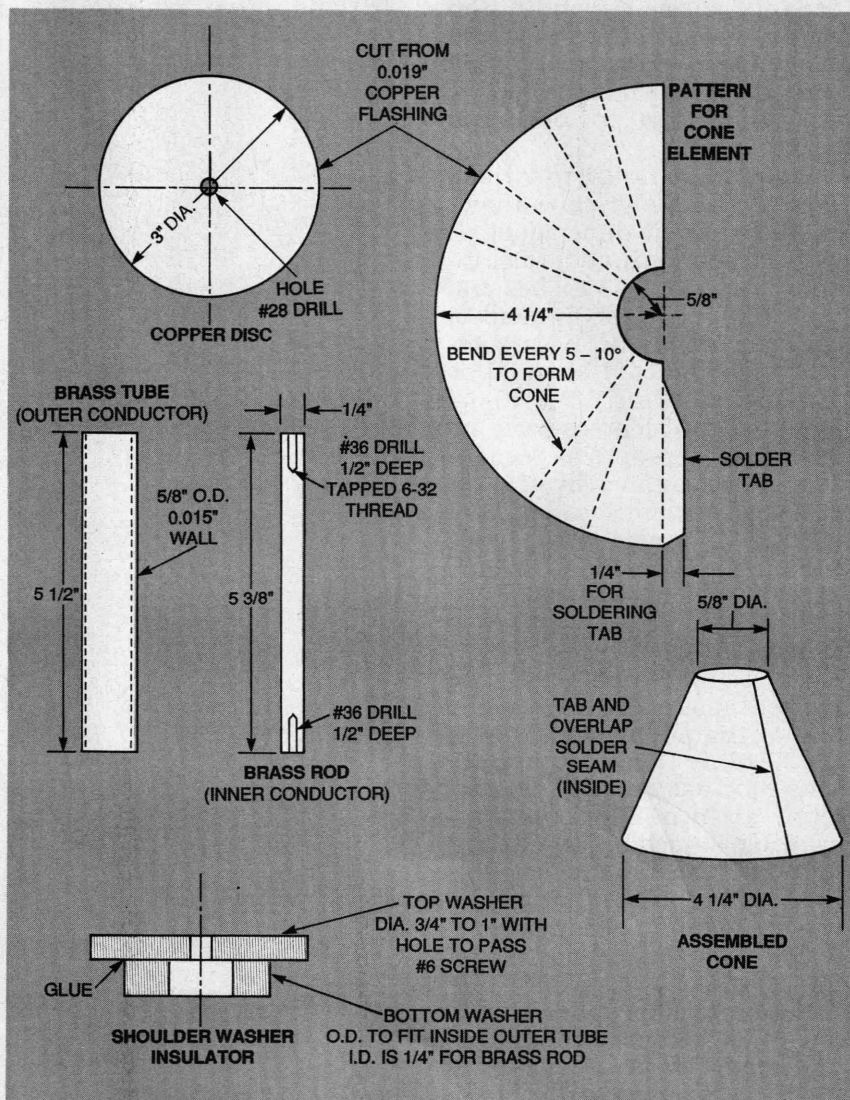


FIG. 2—CONSTRUCTION DETAILS for a 700-2000-MHz UHF discone antenna.

steel wool and solder the cone to the brass tube as shown in Fig. 3. Make sure the brass tube is symmetrical and concentric in the cone. Carefully clean all flux residues using hot water and baking soda, followed by a final rinse in hot water.

Cut the brass rod to the same length as the brass tube. Drill a #36 hole in each end 1/2-inch deep. Use a drill press if possible, and center punch each end to prevent the drill from "walking". The rod has to be held by a vise or clamp to do this. Tap one end of the brass rod for a 6-32 screw thread.

Make a shoulder-washer insulator as shown in Fig. 2 from two plastic washers. The top larger washer should be 3/4-inch diameter by 1/8-inch thick and the center hole should be large enough to pass a 6-32 screw (#28 drill hole), but not larger

PARTS LIST

LIST OF MATERIALS

- 1—Brass tube, 5/8-in. (0.015-in. wall), 5 1/2-in. long
- 1—Brass rod, 1/4-in., 5 3/8-in. long
- 1—Copper or brass sheet, .019 to .030-in. thick, approximately 5 x 12 in.
- 1—Type N UHF connector, UG58A/U, preferably silver plated
- 2—Plastic faucet washers (3/4 to 1-in. dia.) with hole for #6 screw or smaller (drill and file to sizes in drawing)
- 1—6-32 x 1/2-in. brass machine screw, Phillips or slotted head
- 2—Pipe clamps to fit 5/8-inch O.D. tube (plastic preferred) for mounting antenna.

Parts and materials not normally stocked by electronic parts stores can be obtained at hobby shops specializing in model aircraft and/or cars, plumbing supply outlets and hardware stores.

A catalog describing kits for ATV transmitters, ATV receiving converters and other projects usable with the antennas described in this article is available from North Country Radio, PO Box 53, Wykagyl Station, New Rochelle, NY 10804. Please include a #10 SASE and \$1.00 to cover handling and postage.

E-mail: Ncradio200@aol.com
CompuServe 102033,1572

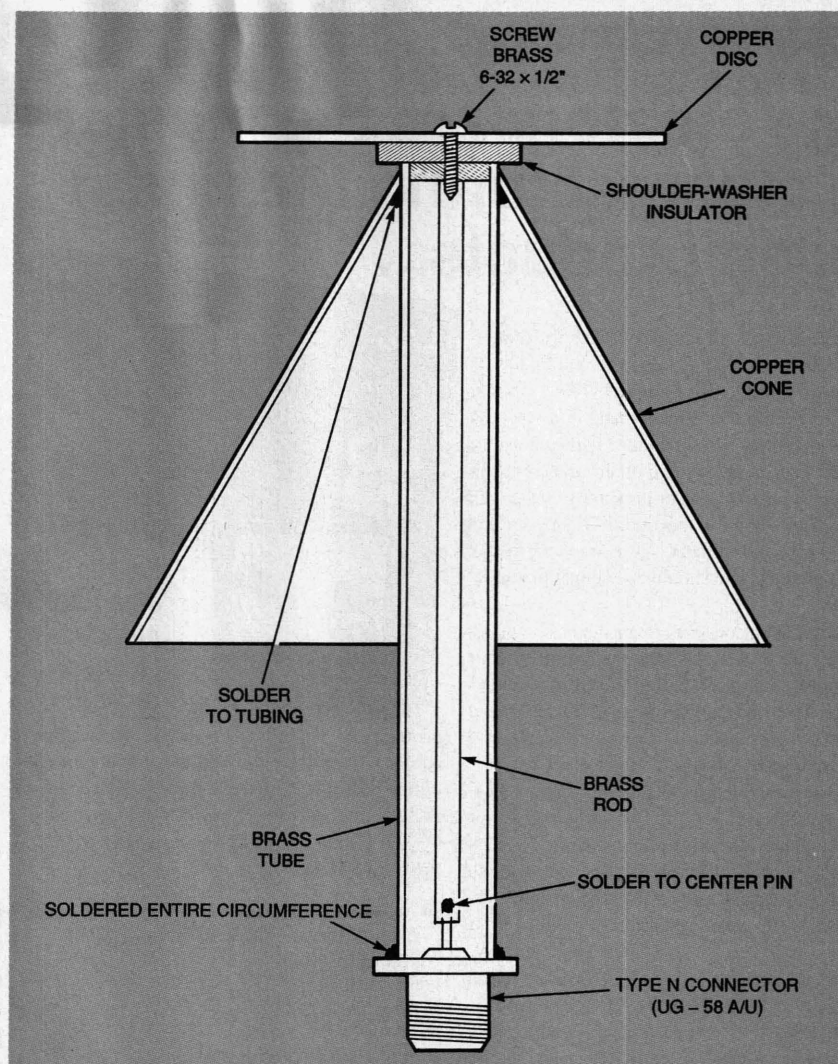


FIG. 3—ASSEMBLY VIEW of the discone antenna. Parts should fit firmly together before soldering is attempted. Clean surfaces to be soldered to a bright shine with #0 steel wool.

than 3/16 inch. The bottom washer should be press fit into the brass tube whose inside diameter is 19/32-inch. The center hole should be 1/4-inch diameter to pass the 1/4-inch brass rod. Glue the washers together to form a shoulder washer. Now trial fit the entire discone assembly together. Trim the length of the center conductor so the top of the shoulder washer rests on the end of the brass tube. When the parts fit properly, you are ready to solder.

Clean the brass rod and the rear of connector flange with fine steel wool. The surfaces should be shiny. Using 60/40 rosin core solder, solder the untapped end of the brass rod to the N type UHF connector's center pin. Use at least a 100-watt

soldering iron. Next, insert this assembly into the lower end of the 5/8-inch brass tube. Insert a 6-32 by 1/2-inch long, brass, roundhead screw through the center of the copper disc, the insulator, and into the tapped hole in the end of the brass rod. Tighten the screw enough to hold the parts together and hold them in place for soldering. Make sure the brass tube is centered on the flange of the connector. Now, solder the connector's flange to the brass tube all around the seam. Use only enough solder to do the job.

Check for shorts with an ohmmeter. There should be an infinite resistance between the disc and cone, and the center terminal and flange of the type N connector. Next, check that zero

Optics and optical experiments

Use lenses, defraction gratings and mirrors to discover new ways to have fun and learn more about lasers and their properties.

THE SINGLE MOST FASCINATING AND USEFUL CHARACTERISTIC OF LASER LIGHT IS PROBABLY COLLIMATION—THE CONCENTRATION OF ITS MONOCHROMATIC LIGHT INTO A TIGHT PENCIL BEAM. FREQUENTLY,

however, this is the very characteristic that needs to be enhanced or defeated to make an experiment or device work. This is where various optical elements come into play. Let's take a look at some of these components individually and examine their capabilities.

Lenses are vital

There are two basic varieties of lenses that are used in laser applications; convex or positive, and concave or negative. They can be either single sided or double sided, (see Fig. 1). Each type has its own special purpose and a particular effect when used as a part of an optical system. Whether single or double, all convex lenses condense light. Concave lenses, on the other hand, spread or diffuse, a beam of light, increasing its area of coverage, but decreasing its intensity.

These properties are important. Especially so, in future columns, when we get to holograms or 3-D pictures. Combinations of both types of lenses are needed to control the light in a variety of experiments. For example, you might start out by using a concave lens to spread the laser beam. Following this a convex lens might be introduced to limit the amount of spread. All in all, an

optical lens is one of the most important optical components you will use with lasers. It is a good idea keep several of each type on hand. The source list provides names and addresses of companies that handle optics. The lenses you buy can be either plastic or glass. Some surplus dealers sell individual elements by the bag. Each bag usually holds a number of lenses of both types.

Surplus lenses are generally seconds, but, a few edge chips, or surface blemishes, rarely affect their performance at for our applications.

WARNING!!!

This article deals with and involves subject matter and the use of materials and substances that may be hazardous to health and life. Do not attempt to implement or use the information contained herein unless you are experienced and skilled with respect to such subject matter, materials and substances. Neither the publisher nor the author make any representations as for the completeness or the accuracy of the information contained herein and disclaim any liability for damages or injuries, whether caused by or arising from the lack of completeness, inaccuracies of the information, misinterpretations of the directions, misapplication of the information or otherwise.

Another good source for lenses is old or damaged cameras. Their lenses can usually be disassembled, and will often provide four or five elements of different styles.

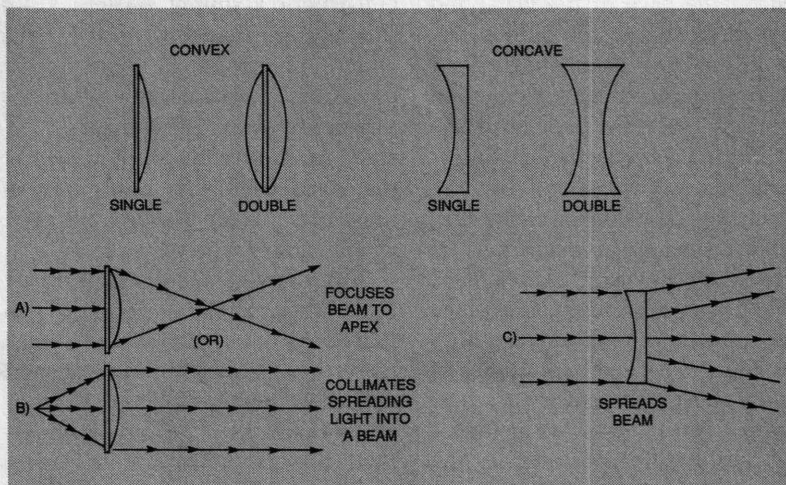


FIG. 1—CONVEX AND CONCAVE LENSES come in both single and double configurations. a—Collimated laser beam into a convex lens produces a focused beam at the focal point of the lens. b—A spreading beam into a convex lens collimates the light. c—Collimated laser beam into a concave lens causes the beam to spread.

Specialized lenses

Fresnels and diffraction gratings are specialized lenses, or more accurately, a composite of specialized lenses. The Fresnel consists of a series of concentric convex elements, eventually terminating in the center. They are readily available in everything from full-page magnifiers, found in the office supply section of discount stores, to laboratory-grade devices. The price is probably the best guide to their quality. Five dollars will usually buy you a good quality specimen for experimenting.

Diffraction gratings are made from either glass or plastic, and are a composite of very small parallel prisms, usually numbering 1500 to 2000 lines per millimeter. When a beam of light strikes their surface it is broken up into a spectrum by each of the mini prisms, which produces a stunning effect. While there are some very valid scientific applications for diffraction gratings, their cost, around \$100 for a 2 inch square, makes them extravagant items for the experimenter. Edmund Scientific sells experimenter-grade replica gratings for a little as \$10.75. Refer to their catalog for details. If available, however, working with them can be a highly entertaining experience.

Mirrors and reflective surfaces

Mirrors come in two varieties; rear surface, where the metallic coating is placed on the back of the glass, and front or first surface, where the coating is on the front of the glass. Rear surface mirrors are the most common as they are the most durable of the two. The metal reflective coating is quite fragile and when it is on the front, is easily scratched or rubbed off.

While ordinary rear-coated mirrors are very useful in many experiments, they do have a major drawback—they are going to reflect two separate laser beams. The metalized back surface of the glass sheet will return the most powerful beam, but a weaker secondary beam will emanate from the front glass of the mirror. Often, this is not a problem, but when it is, the only solution is to use a front-surface mirror that reflects only one image.

Mirrors are easy to find. Ordinary

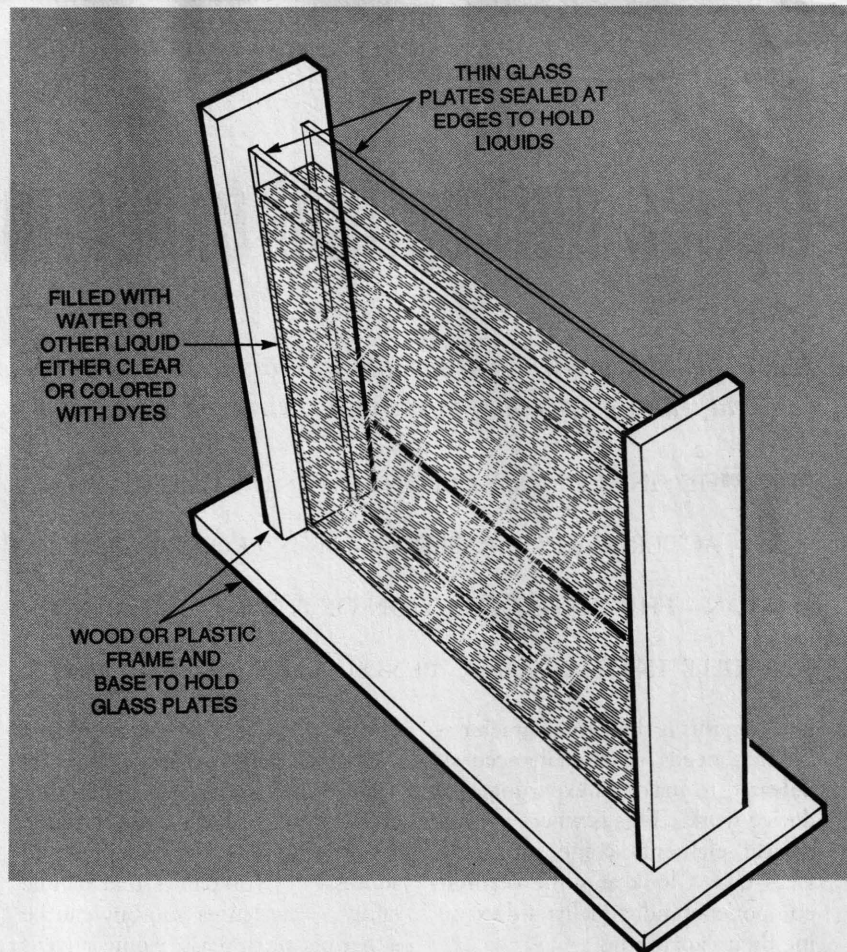


FIG. 2—HOME-MADE LIQUID FILTERS are easy to make. Follow this illustration carefully. Silicon aquarium cement forms a good liquid-proof seal.

rear-surface mirrors are found in discount stores, drug stores, home decorating centers, as well as building and construction supply outlets. The thinner ones are easily cut with a glass cutter. This does take some practice, especially if you have no previous experience working with glass. Go slow, work carefully, and remember that glass edges can result in some nasty cuts. Wear goggles and gloves when working with glass.

Front surface mirrors are harder to find. Many of the companies in the source list, especially surplus dealers, carry them regularly. Hobby shops, toy stores, photographic parts suppliers, old cameras, slide, movie, or overhead projectors can all be fruitful sources of front-surface mirrors. Metalized mylar, and chrome plated or polished sheet metal, while not of exceptional quality, are readily available, inexpensive, and useful substi-

tutes for glass front-surface mirrors.

Once you have the mirrors you need it's time to put them to work. The primary purpose of mirrors and other reflective surfaces is to maneuver the laser beam in a direction other than where it is aimed—to change the direction of the laser beam. In addition to being valuable for the demonstrations and experiments you might set up on an "optical bench", it is easy to see how this principle would be vital in fence perimeter alarms, or bouncing the beam back and forth across a room.

If you add an electric motor to move your mirrors, you'll discover some interesting effects. Try aiming the laser beam first into one revolving mirror and then another. Mirror balls make a really entertaining target.

Prisms and beam splitters

When a beam of white light is

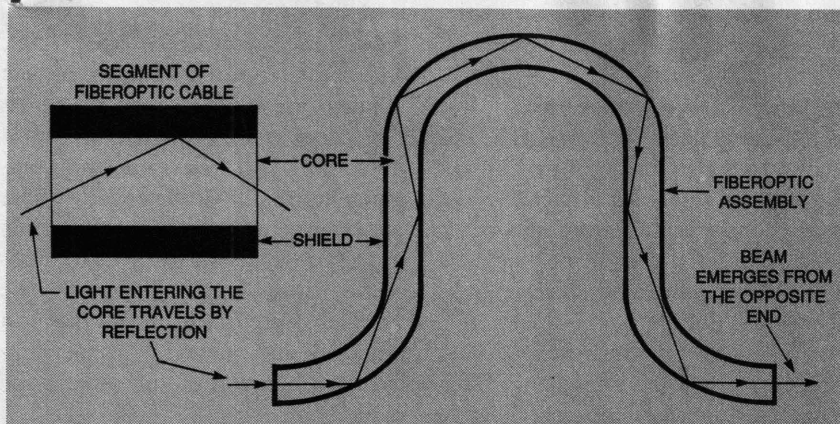


FIG. 3—LIGHT TRAVELS THROUGH A FIBEROPTIC CABLE in a series of straight lines. As each line reaches a surface of the cable it reflects inward and down the cable until the beam finally emerges from the opposite end.

directed at a triangular prism, the light that exits is segmented into all of the individual visible light colors between infrared and ultraviolet. In theory, and by nature, laser light is coherent and monochromatic. In reality, lasers also do emit the entire spectrum, even though one color, or frequency is predominant. Therefore a laser beam directed into a prism will also segment into the full visible light spectrum, with the strongest section being at the spectral location of the primary color. With most He-Ne lasers this will obviously be a section of the red division.

Prisms have many applications when you are conducting optical experiments. Beam splitters are devices that direct part of the light one way, and the rest another. The ratio of this division depends on the type of splitter used and how it is constructed. One simple variety uses the partial mirror principle in which the rear surface is not fully coated. Another method is to place two triangular prisms hypotenuse to hypotenuse. If a right angle or penta-prism is introduced, the beam can be redirected in much the same way as with a mirror, but with virtually no distortion.

The ratio here will depend on the size and shape of the individual prisms. If they are identical the split will be 50 percent each way. If they are different, more light will go in one direction than the other. The most significant benefit of being able to split the beam in two is that you can conduct two different demonstrations or experiments at the same time.

Filters and polarizers

Another optical method for controlling the laser beam calls for using filters. These devices come in a multitude of varieties and purposes. The two most useful filters are single color and white-light inhibitors. The first group are made as red, green, blue, yellow, orange, etc. Generally, they pass only light of their own color and reject all other light. As you saw while building your receivers, this rejection property helps eliminate background noise produced by ambient light. Other dedicated filters control infrared and ultraviolet. Neutral-density have no effect on the color of the light but do regulate the amount of

light that passes through them. Diachronic filters allow only one specific frequency of the spectrum through while rejecting all others. Polarizers or polar screens, are marvels in themselves.

If we align the light waves in only one direction by passing the beam through a polarizer the beam is polarized. When a second polarizing element is placed in front of the first and its polarization is aligned with the first element, both filters are passing light on the same predetermined plane. As long as those planes remain aligned, all of the light will pass through both filters. If you rotate one of the polar screens, the planes fall out of phase and light transmission decreases. When the planes are at 90 degrees to each other, all light is blocked by the second filter. This is theoretical and some tiny fraction of the beam will always pass through, but it is easy to see how effective the polarizer is as a light intensity control. Again, refer to the source list, visit photo suppliers and hobby shops, to locate sources of filters.

Another quite satisfactory resource is clear packaging material or colored gels that come with children's' toys and art kits. They are made of glass, plastic, gel stock. You can make your own from a thin waterproof tank and food coloring dyes. Figure 2 shows all the details of a filter you can make.

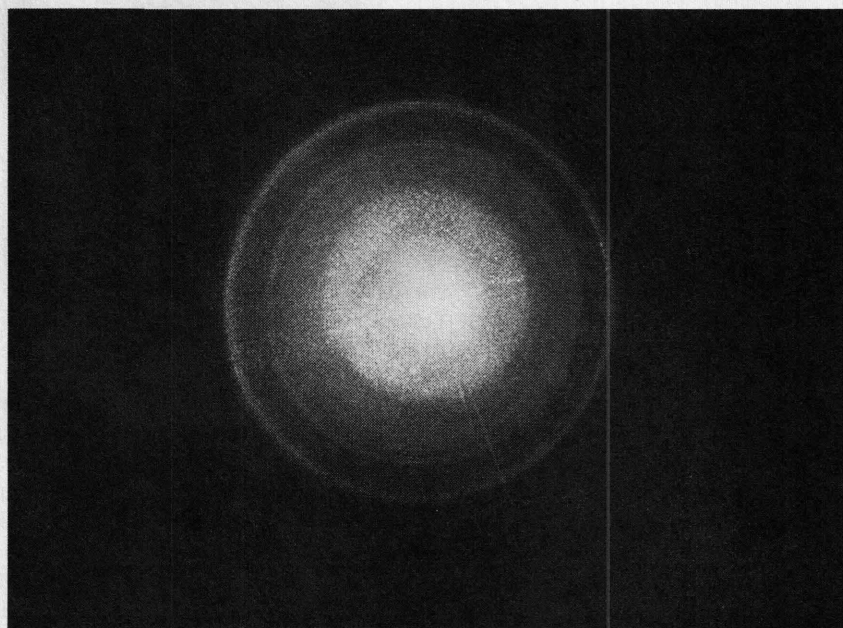


FIG. 4—THIS IS THE HALO EFFECT of laser light. When you duplicate this experiment you will see that each of the bands is a different color.

Light guides and fiber optics

One property of light that all laser experimenters must remember is that light waves or rays or beams do not curve. They only travel in a straight line. This characteristic is the reason why light is easy to redirect. You can change the direction of a laser beam by simply reflecting it off some surface. Also, it contributes to the crisp focusing of light that is not nearly as pronounced in other forms of electromagnetic energy, such as radio or sound. When it is necessary to bend a beam of light, a light guide or fiber optics can handle this task.

Many materials have the ability, through internal reflection, to bend light rays (see Fig. 3). These materials include glass, plastic, and even water. They appear to be function most efficiently when they are formed into round shapes (rods or thin fibers). Again, referring to Fig. 3, the most common type of fiber optic devices uses internal reflection to bounce the

end of the assembly, it will follow the shape of the cable, and emerge from the other end, with very little intensity loss. You can make your own home-brew devices from glass and plastic rods that you wrap with black plastic tape, as a shield to keep the light from escaping from the sides. Both materials can be shaped with the aid of diffused heat, but once set, neither one is flexible.

To demonstrate the principle using water as the light conductor, fill a length of rubber tubing or a piece of ordinary garden hose with water. Then aim the laser beam into one end. The light will pass through the water and emerge from the opposite end.

Another method used to obtain the same result revolves around a core that is, for lack of a better description, "spun" into a composite of convex lenses. In this arrangement, the light is constantly being focused towards the center of the core, and follows the shape in that fashion. While useful,

laser light for use as optical scalpels. New ideas emerge every day.

The communications industry has put them to use in ultra quiet data links that carry, not only voice conversations, but, also, television and computer signals. Commercially, fiber optics have appeared as everything from "gooseneck" flashlights to decorative lamps that shimmer when touched. But, to date, one of the most efficient, and practical partnerships has been between the fiber optic and the laser.

The best bet for finding fiber-optic materials is to check through the source list. Surplus suppliers will be especially valuable. Occasionally, suitable experimental stock can be found in hobby shops or optical or physics kits sold for children and students.

Non-reflective surfaces

A few quick notes regarding the properties of non-reflective surfaces. Any material that tends to absorb, or diffuse the beam is considered non-reflective, and can reveal some unique characteristics of laser light. As previously mentioned, although the laser beam is coherent, it still contains all the other colors in the visible light spectrum, in very small quantities. This can be illustrated by observing the "halo" effect that surrounds the pin point spot of light (see photo in Fig. 4). Note, the circular bands of weak light that ripple out from the center; each is a different color as you will see when you try this experiment.

To set up for this demonstration use a sheet of frosted plastic as the non-reflective surface. Arrange the sheet in a "rear projection" position so the laser points at one side and you are able to look at the other side.

Another interesting property of laser light is speckling. Due to the extreme precision of the beam, and the unevenness of most surfaces, especially those made of non-reflective or semi-reflective materials, what appears to be speckles of light surround the spot (see the photo in Fig. 5). In actuality, this is a sub-reflection of the beam. This type of surface absorbs and stops the light at the end of an optical bench or other experiment or demonstration. It limits the laser beam to its intended area. **EN**

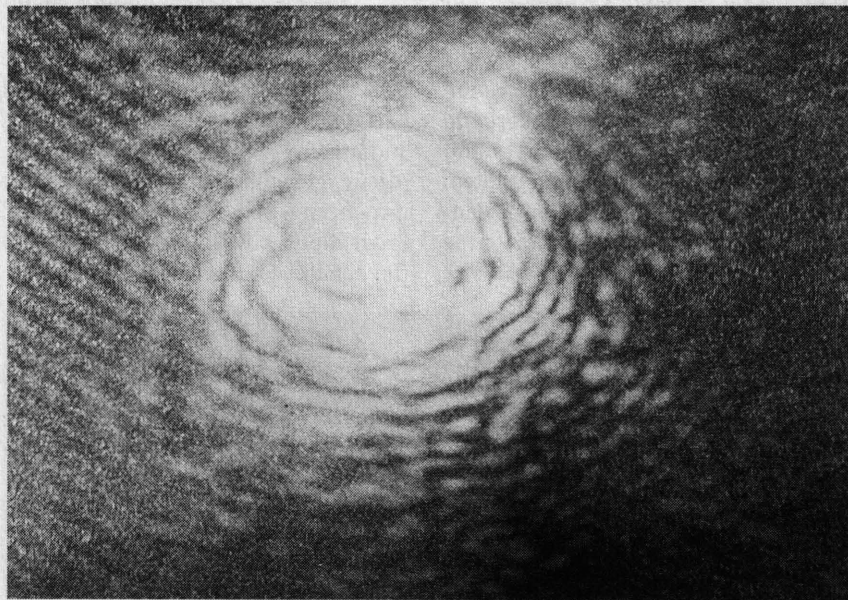


FIG. 5—SPECKLING EFFECT OF LASER LIGHT results when you project the beam on a non-reflective surface. Note the flecks of light around the center. This also reveals the surface terrain of the illuminated object.

light from side to side, down the length of the cable. The solid core, sometimes in actuality a bundle of small fibers, is surrounded by a jacket or shield that keeps external light from entering the fibers. The exposed ends are polished, and sometimes fitted with connectors.

When a laser beam is fed into one

and entertaining on the optical bench, the practical applications of light guides are astounding. They have been used by doctors to illuminate areas hard to reach places, to act as extended lenses to explore internally without surgery, to treat heart and eye diseases when used in conjunction with high-power lasers, that pinpoint

THIS ARTICLE EXPLAINS THE DESIGN of a many different low-power electronic voltage converters. These circuits generate higher-value supply voltages from low-voltage DC sources and negative DC voltages from positive DC voltage sources. The circuits in this article include the popular 555 industry-standard timer IC and the Harris Semiconductor ICL7660 voltage converter IC, devices that simplify the design and construction of a variety of voltage converter circuits.

Low-voltage generation.

The easiest way to increase the voltage level of a DC source or reverse its polarity is to apply that DC voltage to a free-running, squarewave generator whose output is fed to a multisection capacitor-diode voltage-multiplier network. Figure 1 shows block diagrams for positive and negative output voltage converters.

If a positive output voltage is desired, the multiplier must provide a noninverting response, as shown in Fig. 1-a. But if a negative output is required, the multiplier must function as an inverter, as shown in Fig. 1-b.

Voltage converters are based on a variety of multivibrator circuits. These converters can include free-running, square-wave-generating multivibrators based on bipolar or field-effect (FET) transistors or CMOS or TTL integrated circuits. In general, the square-wave generators should oscillate at frequencies between 1 kHz and 10 kHz so that the multiplier section can operate efficiently. This frequency range permits the use of multiplying capacitors with low values.

An easy way to build a voltage converter is to design it around an industry-standard 555 timer IC. (This versatile device and many practical circuits that include it were discussed in the November and December issues of *Electronics Now*, starting on pages 61 and 62, respectively.) The 555 can source or sink up

VOLTAGE

Learn how to build a variety of voltage integrated circuits and put them

to 200 milliamperes when set up as a free-running square-wave generator.

Figures 2, 3, 4, and 5 are voltage-converter circuits based on the 555. The 555 is configured

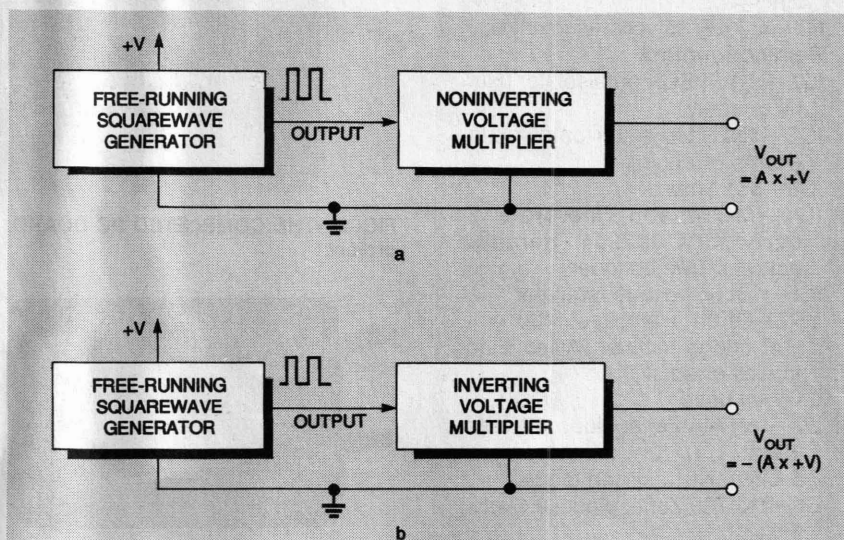


FIG. 1—VOLTAGE CONVERTER BLOCK DIAGRAMS: positive output (a) and negative output (b).

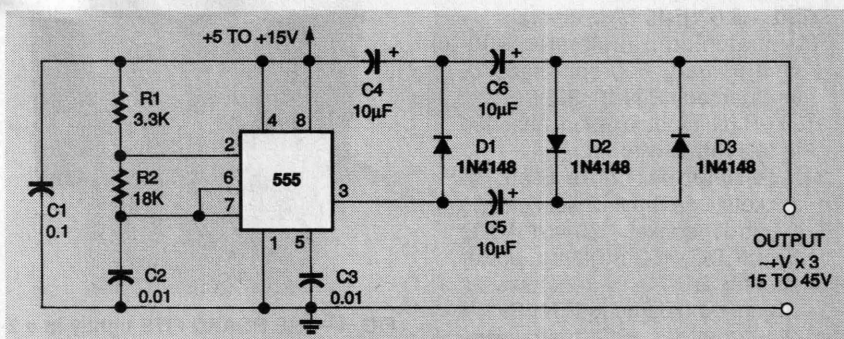


FIG. 2—DC VOLTAGE-DOUBLER circuit based on the 555 timer IC.

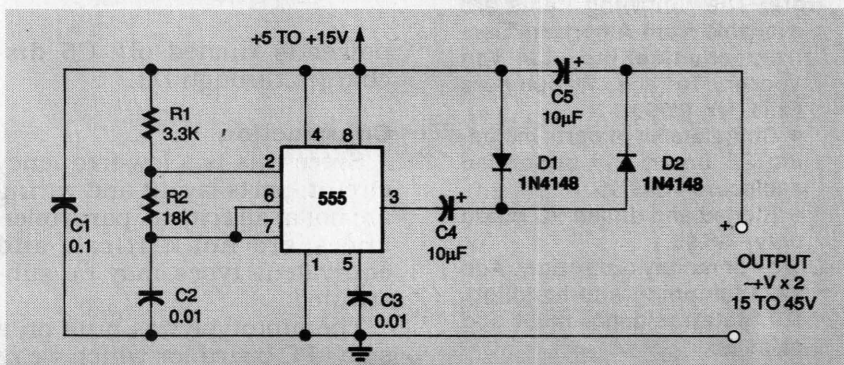


FIG. 3—DC VOLTAGE-TRIPLER circuit based on the 555 timer.

CONVERTERS

-conversion circuits from popular to work in your own circuit designs.

as a free-running astable multivibrator (squarewave generator) operating at about 3 kHz in all of these circuits. The 3-kHz frequency is determined by the values of resistors R1 and R2 in series with capacitor C2. Supply-line capacitor C1 prevents the 3-kHz output of the 555 from feeding back into the device, and capacitor C3 enhances circuit stability.

The circuit in Fig. 2 is a DC voltage-doubler that generates a DC output voltage with a value that is approximately twice that of the supply voltage. The 555's output is fed to a voltage-doubler network made up of capacitors C4 and C5 and diodes D1 and D2. The network produces the approximate *two times supply voltage output* when that output is unloaded.

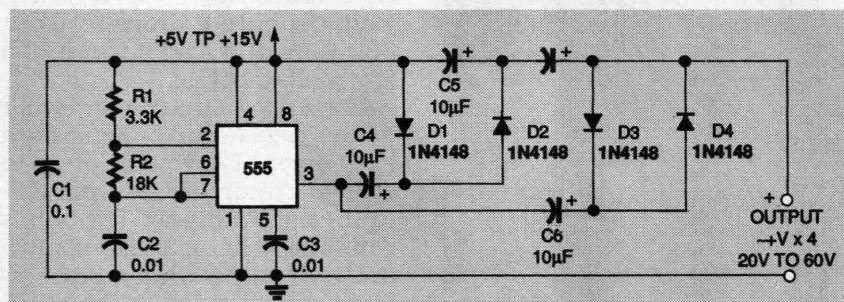


FIG. 4—DC VOLTAGE-QUADRUPLER circuit based on the 555 timer.

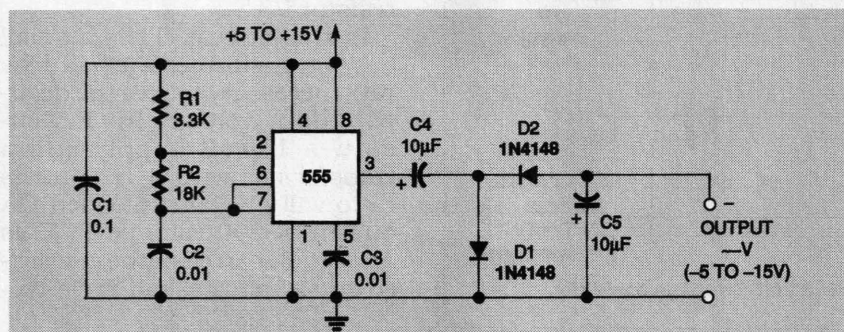


FIG. 5—DC NEGATIVE VOLTAGE generator based on the 555 timer.

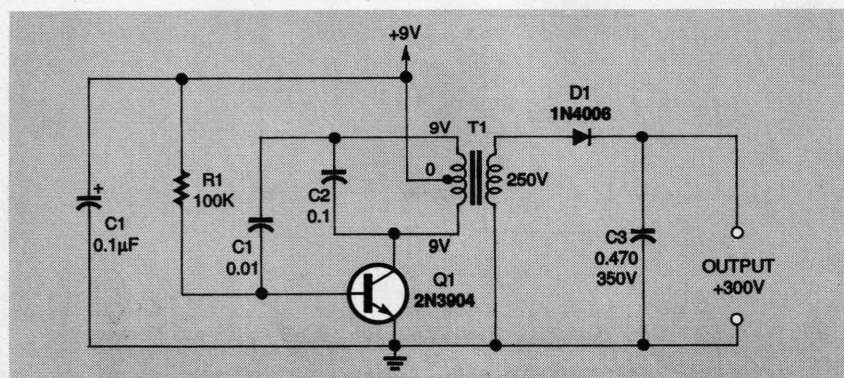


FIG. 6—NINE-VOLT-TO-300-VOLT DC-to-DC converter.

The precise output value is:

$$2 \times V_{\text{peak}} - (V_{\text{fd1}} + V_{\text{fd2}})$$

Where: V_{peak} is the peak output voltage of the squarewave generator V_{fd1} and V_{fd2} are the forward voltage drops (about 600 millivolts) of multiplier diodes D1 and D2.

The output voltage of the circuit decreases when the output is loaded.

The Fig. 2 voltage doubler circuit will work with any DC supply of 5 to 15 volts. Because of its voltage-doubling response, it can provide voltage outputs over a range of about 10 to 30 volts. Higher voltage can be obtained by adding more multiplier stages to the circuit.

The Fig. 3 DC voltage tripler circuit can provide output voltages from 15 to 45 volts. The Fig. 4 DC voltage quadrupler circuit can provide output voltages from 20 to 60 volts.

Figure 5 is the schematic for a DC negative-voltage generator, capable of producing an output voltage that is approximately equal in amplitude but opposite in polarity to that of the power supply. It also operates at 3 Hz, and it drives an output stage consisting of capacitors C4 and C5 and diodes D1 and D2. This circuit's *split-supply* output is useful for powering ICs that require both positive and negative power from a single-ended source.

High-voltage generation.

The generation of higher output voltages with voltage multipliers is usually cost-effective only when the values of the required voltages are less than six times the supply voltage. Where very large step-up ratios are required (e.g., hundreds of volts generated from a 12-volt supply), other circuit designs are recommended.

It is usually cost-effective to apply the output of the low-voltage oscillator or squarewave generator as the input to a suitable step-up voltage transformer. The required high AC output voltages appear across the transformer's secondary winding.

This AC voltage can easily be converted back to DC with a

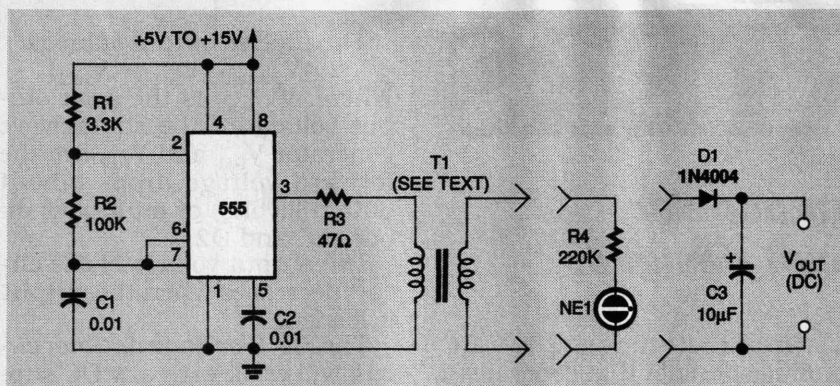


FIG. 7—HIGH-VOLTAGE GENERATOR that can also be a neon lamp driver.

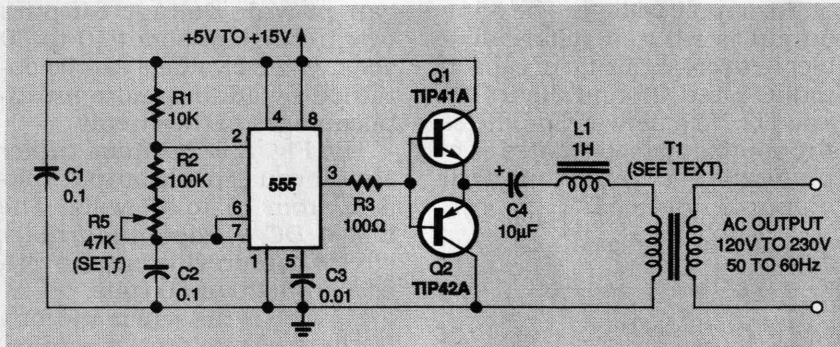


FIG. 8—DC-TO-AC INVERTER based on the 555 IC.

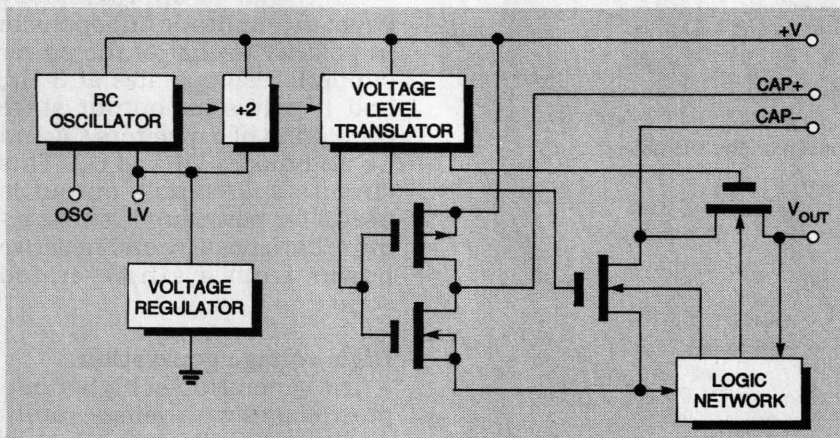


FIG. 9—FUNCTIONAL DIAGRAM for the ICL7660 CMOS voltage converter.

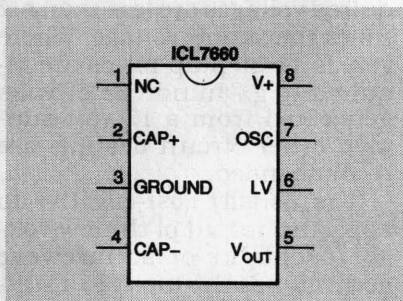


FIG. 10—PIN CONFIGURATION for an ICL7660 in an eight-pin miniDIP.

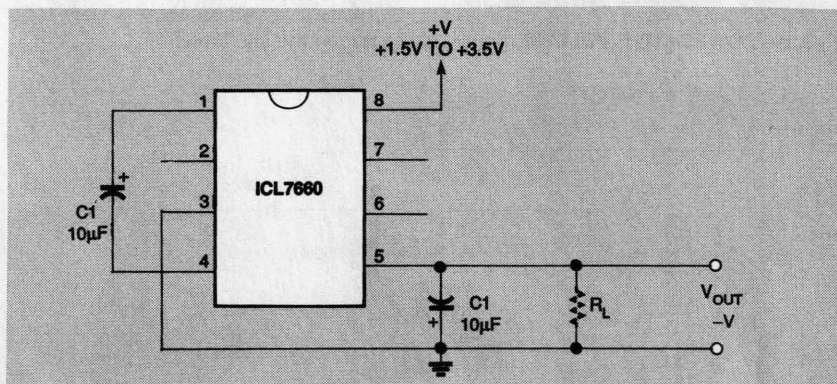


FIG. 11—NEGATIVE VOLTAGE CONVERTER for a 1.5- to 3.5-volt supply.

Figure 6 is a DC-to-DC converter circuit that can generate a 300-volt DC output from a 9-volt DC power source. In this circuit, transistor Q1 and its associated circuitry form a Hartley inductive-capacitive (LC) oscillator. The voltage on the primary winding that swings between zero and 9-volts is applied to 250-volt transformer T1. The inductance of the primary is the inductive element in the LC oscillator, which is tuned by capacitor C2.

The supply voltage is stepped up to about 350 volts peak at the secondary of T1. This output is half-wave rectified by diode D1, and it charges capacitor C3. Without a permanent load on C3, the capacitor can deliver a powerful but non-lethal electrical shock. With a permanent load, the output drops to about 300 volts at a load current of a few milliamperes.

The Fig. 7 circuit can either drive a neon lamp or generate a low-current DC voltage of up to several hundred volts from a 5- to 15-volt DC supply. The 555 is configured as a 3-kHz astable multivibrator whose square-wave output is fed to the input of transformer T1 through the resistor R3.

In this circuit, T1 is a small audio transformer with a turns ratio necessary to give the desired output voltage. In this circuit, a 10-volt supply and a transformer with a 1:20 turns ratio will give an unloaded DC output of 200 volts peak. This AC voltage can easily be converted to DC with a half-wave rec-

simple rectifier-filter network, if required. Figures 6, 7 and 8 show schematics for low-power,

high-voltage generation circuits based on of this principle.

tifier and filter capacitor, as shown in Fig. 7.

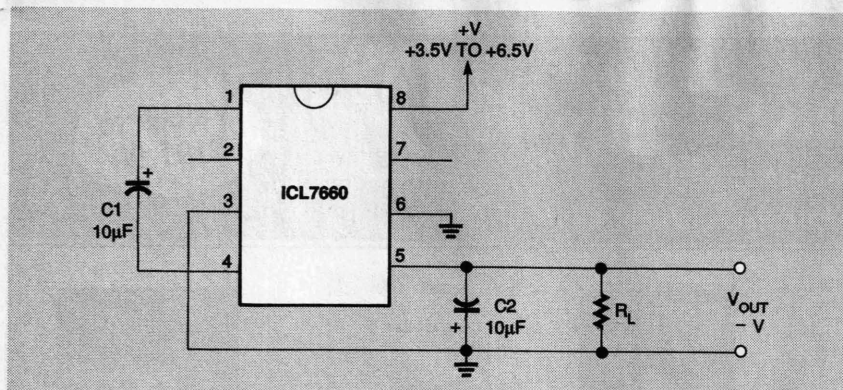


FIG. 12—NEGATIVE VOLTAGE CONVERTER for a 3.5- to 6.5-volt supply.

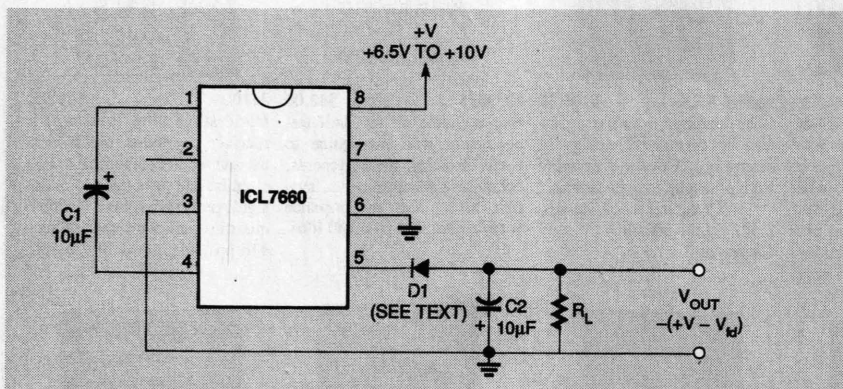


FIG. 13—NEGATIVE VOLTAGE CONVERTER for a 6.5- to 10-volt supply.

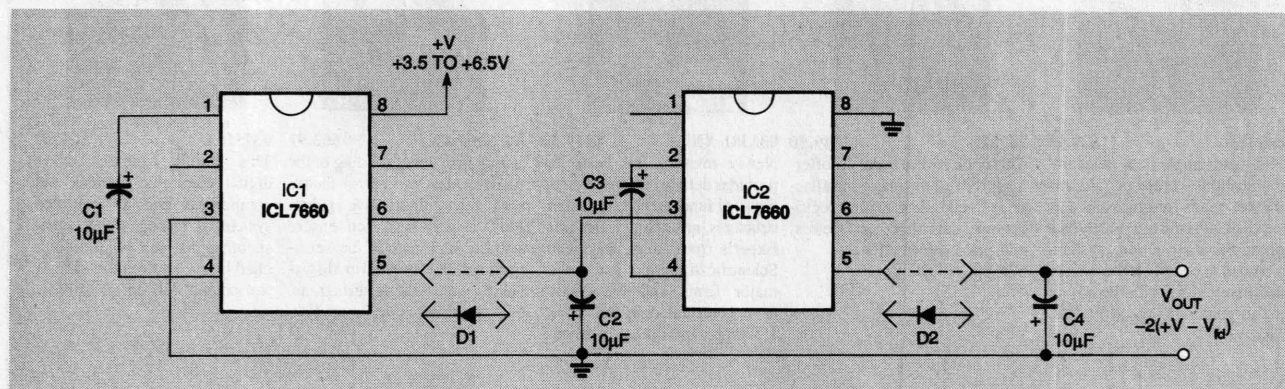


FIG. 14—METHOD FOR CASCADING voltage converter ICs for increased voltage.

The DC-to-AC inverter circuit in Fig. 8 will produce an AC output at line frequency and voltage. The 555 is configured as an astable multivibrator whose output frequency can be adjusted over a range of 50 to 60 Hz by trimmer potentiometer R5.

The output of the 555 is fed to the paired NPN and PNP bipolar transistors Q1 and Q2, and their output is fed into the low-voltage side of reverse-connected filament transformer T1. It has the necessary step-up turns ratio. Capacitor C4 and coil L1 filter the input to ensure

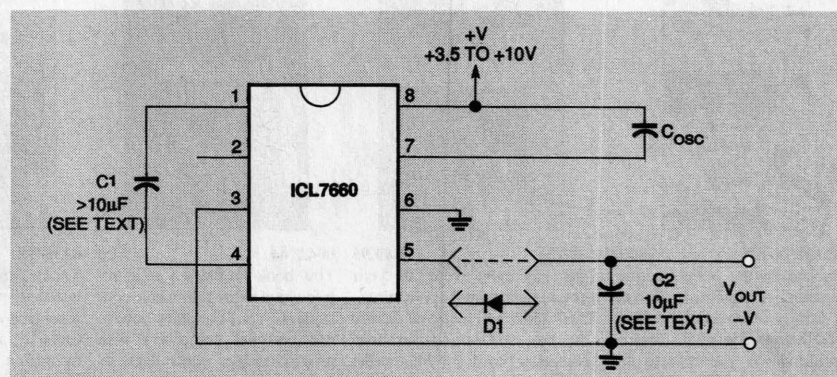


FIG. 15—CAPACITOR SUBSTITUTION for reducing the voltage converter's oscillator frequency.

that the transformer input is essentially a sine wave.

CMOS voltage converter

The Harris ICL7660 is a monolithic CMOS voltage converter that converts its supply voltage from positive to negative for an input range of +1.5 volts to +10 volts. This results in complementary output voltages of -1.5 volts to -10 volts. For example, if powered from a +5-volt source, the device will generate a -5-volt output.

The ICL7660 has a typical open-circuit voltage conversion efficiency of 99.9% and a typical power efficiency of 98%. When the output is loaded, the ICL7660 acts like a voltage source with an output impedance of about 70 ohms. It can supply maximum currents of about 500 microamperes.

Only two external capacitors are needed to perform the charge pump and charge reservoir functions. The ICL7660 can also be configured as a voltage doubler and it will generate output voltages up to +18.6 volts with a +10-volt input.

Figure 9 is the functional block diagram of the ICL7660.

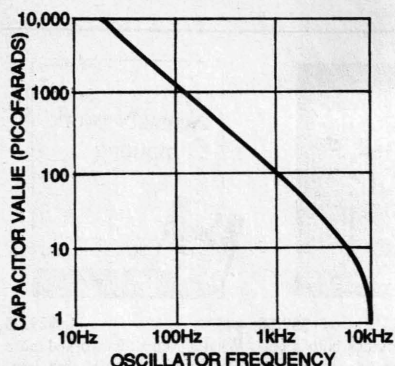


FIG. 16—FREQUENCY OF OSCILLATION as a function of external oscillator capacitor value (+V = 5 volts).

It contains an RC oscillator, divide-by-2 circuit, voltage-level translator, voltage regulator, four output power MOS switches, and an unusual logic network. The logic network senses the most negative voltage in the device and ensures that the output N-channel switch source-substrate junctions are not forward biased. This is intended to ensure latchup-free operation.

The unloaded oscillator oscillates at a nominal frequency of 10 kHz when the input supply voltage is 5.0 volts. This fre-

quency can be lowered by the addition of an external capacitor to the osc terminal, or the oscillator can be overdriven by an external clock.

The LV terminal can be tied to GROUND to bypass the internal series regulator and improve low voltage (LV) operation. At medium to high voltages (+3.5 to +10.0 volts), the LV pin is left floating to prevent device latchup.

The ICL7660 is available in eight-pin miniDIP and small-outline IC (SOIC) packages as well as TO-99 cases. Figure 10 is the drawing of pin configurations for the eight-pin miniDIP package, the least expensive version. Power dissipation of this package is 300 milliwatts. The ICL7660S, an enhanced direct replacement for the ICL7660, is now available and can be substituted if desired.

The operation of the ICL7660 is similar to that of the oscillator and voltage-multiplier shown in Fig. 2, but it performs that function more efficiently. As stated earlier, the use of conventional silicon multiplier diodes causes the circuit's unloaded output voltage to drop by about 1.2 volts—the sum of the forward voltage drops of the two diodes.

The ICL7660 eliminates the voltage drop by replacing the diodes with MOS power switches. These switches are driven by the logic network so that each switch automatically closes when it is forward biased and opens when it is reverse biased. This feature accounts for its high operating efficiency.

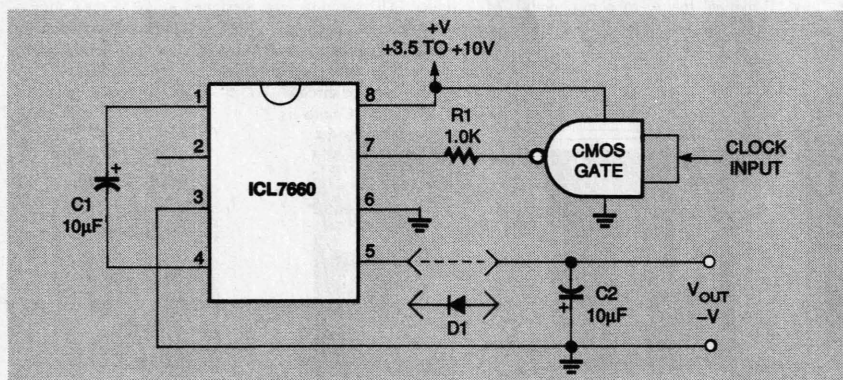


FIG. 17—EXTERNAL CLOCKING METHOD for reducing the internal oscillator frequency of the ICL7660.

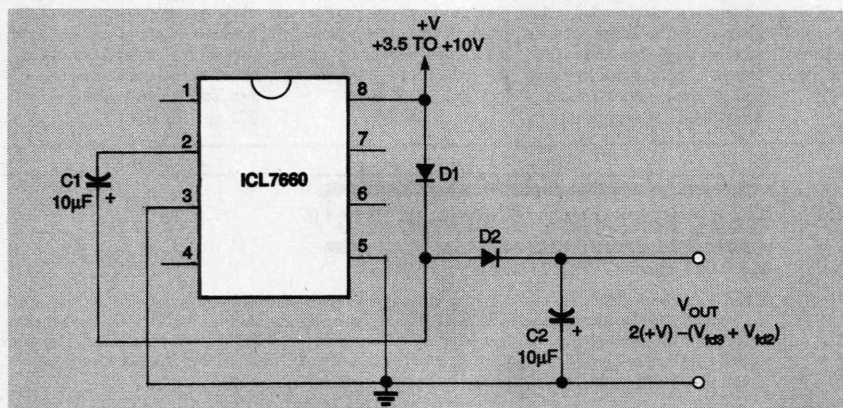


FIG. 18—POSITIVE VOLTAGE MULTIPLIER based on the ICL7660.

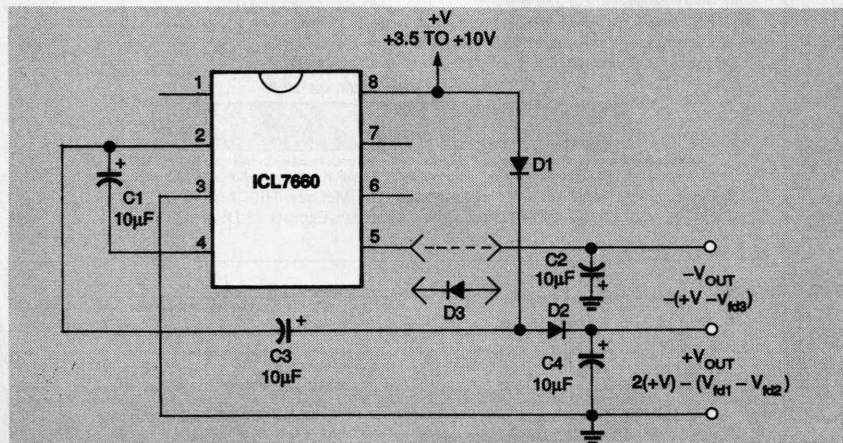


FIG. 19—COMBINED NEGATIVE VOLTAGE converter and positive doubler.

Do's and don'ts

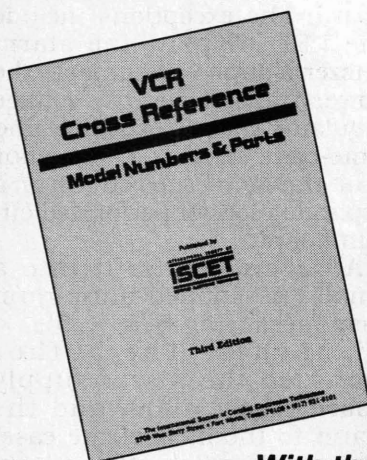
The manufacturer recommends some *Do's and Don't's* to assure reliable device operation:

1. Do not exceed the device's maximum supply voltage.
2. Do not connect any pins to voltages greater than +V or less than ground.
3. If the IC is operating in the power supply range of 1.5 to 3.5 volts, ground the LV-SX pin 6; at supply values greater than 3.5 volts, leave pin 6 open circuited.
4. Do not short circuit the output to the +V supply for volt-

Continued on page 62

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VOLTAGE CONVERTERS

continued from page 50

ages above 5.5 volts for extended periods.

5. When installing polarized capacitors, connect the positive lead of the capacitor to the CAP + pin 2 and the negative lead to the CAP- pin 4 of the ICL7660. Connect the positive terminal of capacitors to V_{OUT} pin 5 to ground.

6. Be sure that pin 5 does not swing more positive than GND pin 3 to prevent latchup

7. If the supply voltage is greater than 6.5 volts, put a protective diode in series with pin 5.

Practical ICL7660 Circuits.

The most popular application for the ICL7660 is as a negative voltage converter. Figures 11, 12, and 13 are three basic negative converter circuits. In each circuit, capacitors C1 and C2 are *multiplier* capacitors with a value of 10 microfarads.

The negative voltage converter circuit in Fig. 11 is designed to be powered from a 1.5- to 3.5-volt supply, and it requires only two external electrolytic capacitors, C1 and C2. The circuit in Fig. 12 is similar, but it is designed to be powered in the 3.5- to 6.5-volt range and its LV pin 6 grounded.

The negative voltage converter circuit in Fig. 13 is designed for a 6.5- to 10-volt supply. Diode D1 is in series with +V_{OUT} pin 5 to protect the IC from excessive reverse biasing by C2 when the power supply is removed. Diode D1, which should be a Schottky diode, reduces the available output voltage by V_{fd}.

Up to ten cascading ICL7660s will give increased output voltage. For example, if three ICs are cascaded, their final output voltage will be -3 volts. Figure 14 shows how to make the connections for cascading two of these stages. Connect all additional stages in the same way as shown in the schematic.

Output frequency change

In some applications there might be a reason for reducing the operating frequency of the

IC's RC oscillator. This can be done by inserting an external oscillator capacitor C_{OSC} between osc pin 7 and +V_{pin} 8, as shown in Fig. 15.

The relationship between the value of C_{OSC} in picofarads and the oscillator frequency f_{OSC} is shown in Fig. 16. For example, a C_{OSC} of 100 pF reduces the operating frequency by a factor of ten, from 10 kHz to 1 kHz.

However, it will be necessary to alter the values of capacitors C1 and C2 to compensate for this 10 to 1 reduction of frequency if circuit efficiency is to be kept high. Increase the values of C1 and C2 (both 10 microfarads) by the inverse of the reduction function. For example, if the oscillator frequency is to be 1 kHz, both C1 and C2 should have values of about 100 microfarads.

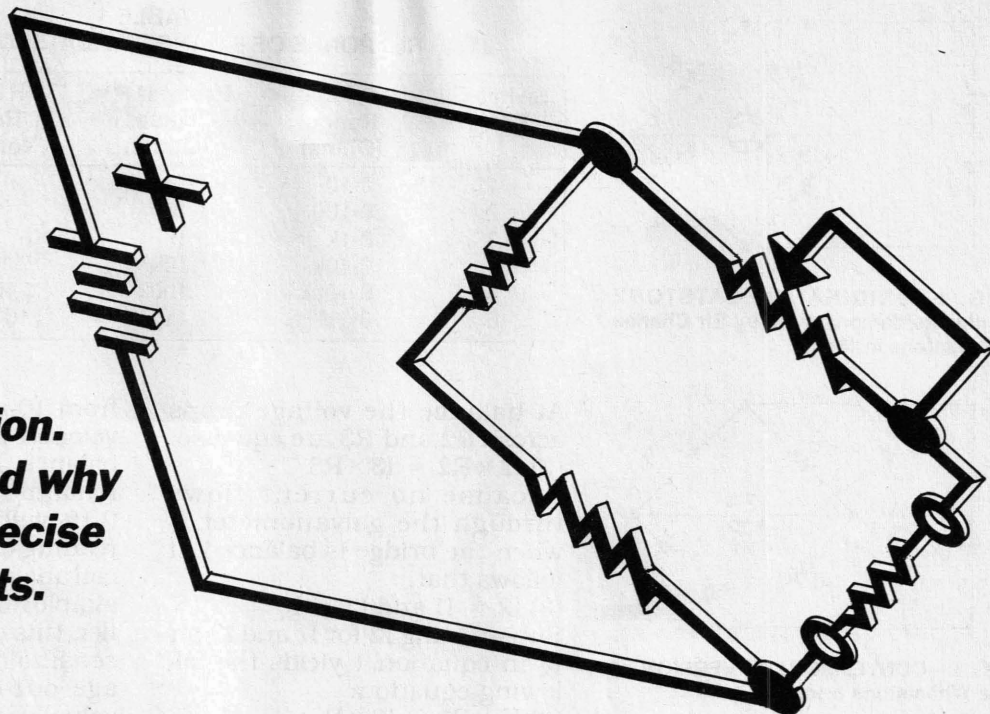
Oscillator frequency can also be reduced by connecting osc pin 7 as shown in Fig. 17 so that an external clock input overdrives the internal oscillator. Feed the clock signal to pin 7 through 1-kilohm series resistor R1. The clock signal should switch fully between the two power supply values. The CMOS NAND gate is connected as an inverting buffer stage to ensure reliable switching between those values.

The ICL7660 can also be a positive voltage multiplier. It can produce a positive output voltage that is nearly double the initial supply voltage value when set up as shown in Fig. 18. The oscillator output signal at CAP+ pin 2 drives a capacitor-diode voltage-doubler network. It is the same network as the one shown in Fig. 2.

The two diodes, D1 and D2, reduce the available output voltage by an amount equal to their combined forward volt drops. Consequently, both D1 and D2 should be low-loss germanium diodes.

Figure 19 shows how the circuits of Figs. 12, 13, and 18 can be combined to form a positive voltage multiplier and negative voltage converter that provides dual output voltages. Each voltage source has an output impedance of about 100 ohms. Ω

The bridge circuit holds an honored place in the history of electronic instrumentation. Learn how and why they make precise measurements.



BRIDGE CIRCUITS

RAY MARSTON

A BRIDGE CIRCUIT IS AN INSTRUMENT for making *comparison measurements*. These circuits are widely used to measure resistance, inductance, capacitance, and impedance. They have four sections or arms connected in series to form a diamond. An AC or DC voltage source is connected between one pair of opposite junctions, and an indicating meter or output circuit is connected between the other pair of opposite junctions.

Bridge circuits operate on a *null-indication* principle. When the bridge is balanced, the output is zero. Consequently, the indication is independent of the calibration of the indicating device. For this reason, accurate measurements can be made with bridges. Bridge circuits can also control other circuits. When functioning as controls, one arm of the bridge contains a resistive element that is sensitive to such physical variables as temperature or pressure.

Bridge circuits are classical

electrical measuring instruments predating electronics as a field distinguished from classical electrical engineering. Until about 25 years ago, bridge circuits were the only practical instruments for making many kinds of precise measurements.

However, in many applications they have been displaced by digital multimeters that contain provisions for making comparable measurements. Nevertheless, the availability of low-cost digital multimeters and capacitance meters (sometimes combined in one instrument) does not negate the value of the basic bridge circuit.

The Wheatstone bridge

The Wheatstone bridge is the best known bridge circuit and the one most widely taught in electronics courses. It consists of two parallel resistance branches with each branch containing two series elements, typically resistors. A DC voltage source is connected across a diamond-shaped network to provide a source of current through

the network. A *null detector*, originally a simple galvanometer, is connected between the parallel branches to detect a condition of balance.

The circuit shown in Fig. 1 was first invented by S.H. Christie in 1833. However, its value was not recognized until 1847 when Sir Charles Wheatstone described how it could make accurate electrical measurements. As a result of his insights and demonstrations, the bridge came to be known as the Wheatstone bridge.

The Wheatstone bridge has probably been in use longer than any other electrical measuring instrument. Although still an accurate and reliable instrument, it is not as convenient to use as the new digital multimeters. Nevertheless, 0.1% accuracy is commonly obtainable with the Wheatstone bridge. This compares with the 3% to 5% error expected in resistance measurements made with analog ohmmeters. Accuracy on digital multimeters will depend on the meter's

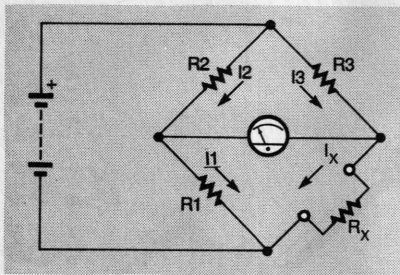


FIG. 1—ORIGINAL WHEATSTONE bridge as demonstrated by Sir Charles Wheatstone in 1843.

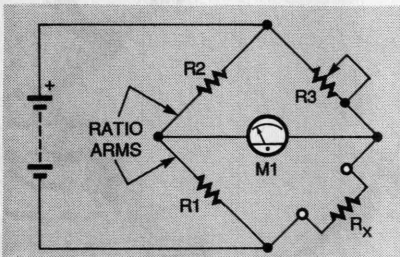


FIG. 2—CONTEMPORARY VERSION of the Wheatstone bridge.

characteristics, such as DC voltage accuracy and resolution.

When making measurements with a Wheatstone bridge to determine the value of an unknown resistor (R_x) one of the remaining resistors is varied until the current through the null detector decreases to zero. The bridge is then in a balanced condition. This means that the voltage across resistor R_3 is equal to the voltage drop across the unknown resistor R_x and the two divider resistors R_1 and R_2 are passing equal voltages. As a result, the following is true: (1) $I_1 \times R_1 = I_x \times R_x$

At balance the voltage drops across R_2 and R_3 are equal so:

$$(2) I_2 \times R_2 = I_3 \times R_3$$

Because no current flows through the galvanometer G when the bridge is balanced, it follows that:

$$(3) I_2 = I_1 \text{ and } I_3 = I_x$$

Substituting I_2 for I_1 and I_3 for I_x in equation 1 yields the following equation:

$$(4) I_2 \times R_1 = I_3 \times R_x$$

Dividing equation 2 by equation 4, yields:

$$(5) R_2/R_1 = R_3/R_x$$

This can be rewritten as:

$$(6) R_2 \times R_x = R_3 \times R_1$$

Equation 6 states the conditions of balance of a Wheatstone bridge and is useful for computing the value of an unknown resistor, once balance has been achieved.

Dividing both sides by R_1 yields the most common Wheatstone bridge equation:

$$R_x = R_3 \times R_1/R_2$$

An important feature of the original Wheatstone bridge was its very high null sensitivity. Thus, if the bridge is energized

from 10-volt DC, 5 volts is developed across all resistors at balance, and the galvanometer remains centered. A shift of only 0.1% will then give a 5 millivolt reading on a calibrated moving-coil analog meter. By installing a simple null-detecting DC amplifier, this circuit can have a *null sensitivity* factor (i.e., percentage out-of-balance detection value) of about 0.003%.

However, the major drawback of this 1847 bridge is that R_3 must have a large resistive range if it is to balance all possible values of R_x . In 1848, the German-born engineer Wilhelm von Siemens, overcame this drawback by introducing the modifications, as shown in Fig. 2

The bridge can be further modified as shown in Fig. 3 by giving R_2 a fixed value and making R_1 switch-selectable. This version is based on a 1970 version, high-accuracy laboratory measuring instrument.

The circuit in Fig. 3 can measure DC resistances from near-zero to 1 megohm in six switch-selected decade ranges. Resistor R_3 is a calibrated 10-kilohm variable potentiometer that controls the sensitivity of the balance-detecting center-zero meter. The value of R_L limits bridge current to a few milliamperes. Table 1 lists the major weakness of this 1970 Wheatstone bridge—its null sensitivity (which is proportional to the R_x test voltage) degrades in proportion to the R_1/R_2 ratio's divergence from unity.

Consequently, the sensitivity is nominally 0.003% on the 10-kilohm range where the R_1/R_2 ratio is 1/1, but it degrades to

Continued on page 145

TABLE 1
RESPONSE OF 6-RANGE WHEATSTONE BRIDGE

Switch S1 Range	Bridge Range (Ohms)	Resistor R1 Value (Ohms)	R1/R2 Ratio (Nominal)	Bridge Null Sensitivity
1	0-10	10	1/1000	3.0 %
2	0-100	100	1/100	0.3 %
3	0-1k	1k	1/10	0.03 %
4	0-10k	10k	1/1	0.003 %
5	0-100k	100k	10/1	0.03 %
6	0-1MEG	1MEG	100/1	0.3 %

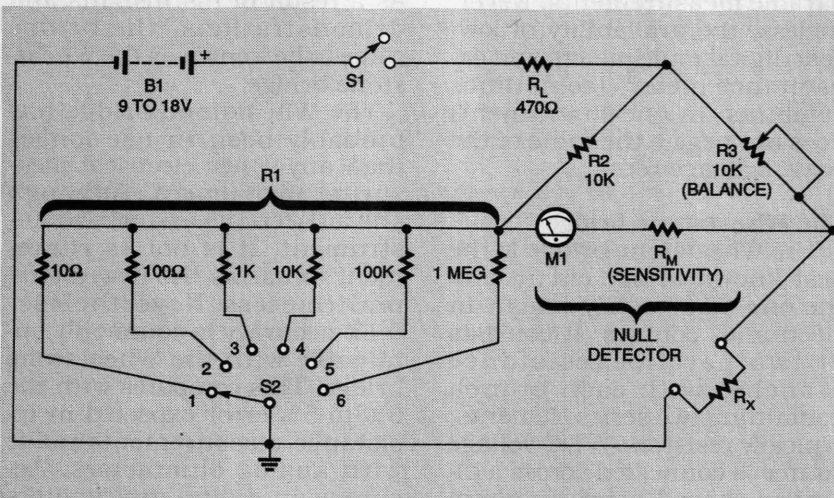


FIG. 3—WHEATSTONE BRIDGE with six DC measuring ranges.

BRIDGE CIRCUITS

continued from page 42

0.3% on the 100-ohm and 1-megohm ranges where the $R1/R2$ ratios are 1/100 and 100/0, respectively. To be a practical instrument, the Fig. 3 circuit must include a sensitive null-balance detector.

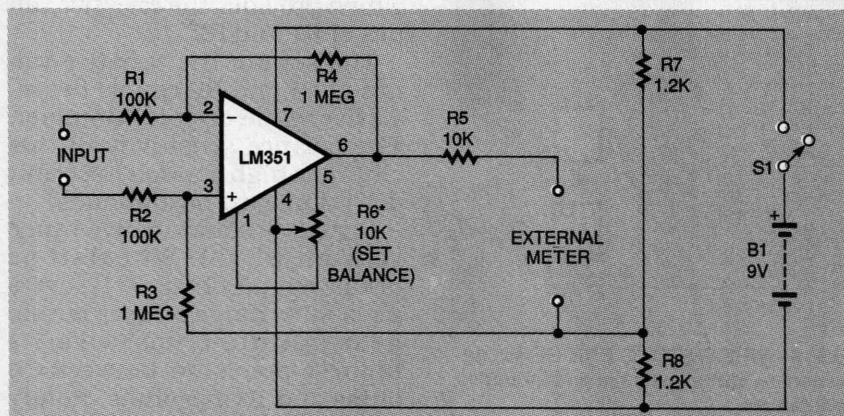


FIG. 4—A DC NULL POINT AMPLIFIER for use with an external digital multimeter or analog volt-ohmmeter

Figure 4 is a schematic for a $\times 10$ DC differential amplifier for use with an external analog volt-ohmmeter to form such a detector. This circuit must have its own independent 9-volt power supply. The LF351 is a low-cost BIFET operational amplifier available from Motorola, Texas Instruments, National Semiconductor, and others.

ampere or 100-microampere range for high-sensitivity. In the 100-microampere range, the circuit must first be balanced by short-circuiting its input terminals together and trimming the multiturn 10-kilohm SET-BALANCE control for a zero reading on the meter.

Wheatstone variations

The Wheatstone bridge circuit of Fig. 2 can be configured in three other ways without invalidating the basic balance equations, as shown in Figure 5. In each form $R1/R2$ are known as the bridge's ratio arms. Notice that the bridge's signal-source and detector terminals can be transposed without upsetting the circuit's balance equations. This also holds true for the various versions of capacitance and inductance bridges. The most useful Wheatstone bridge variation is that shown in Fig. 5-a.

Figure 6 shows a modern six-

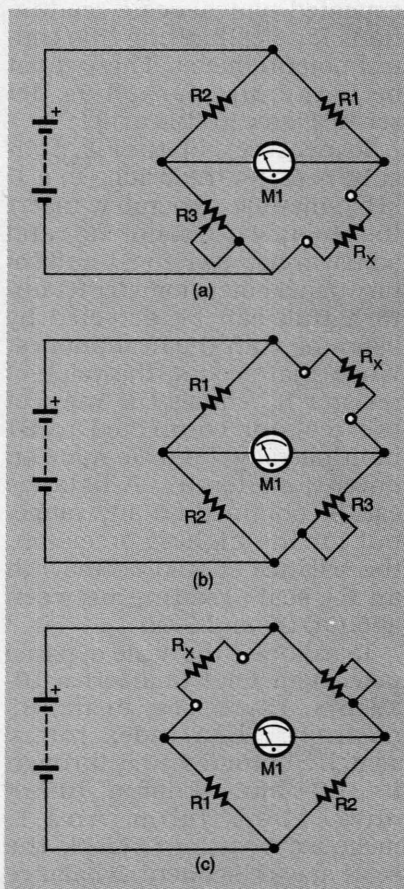


FIG. 5—THREE ALTERNATIVE VERSIONS of Wheatstone bridges follow the same rules as the circuit in Fig. 2.

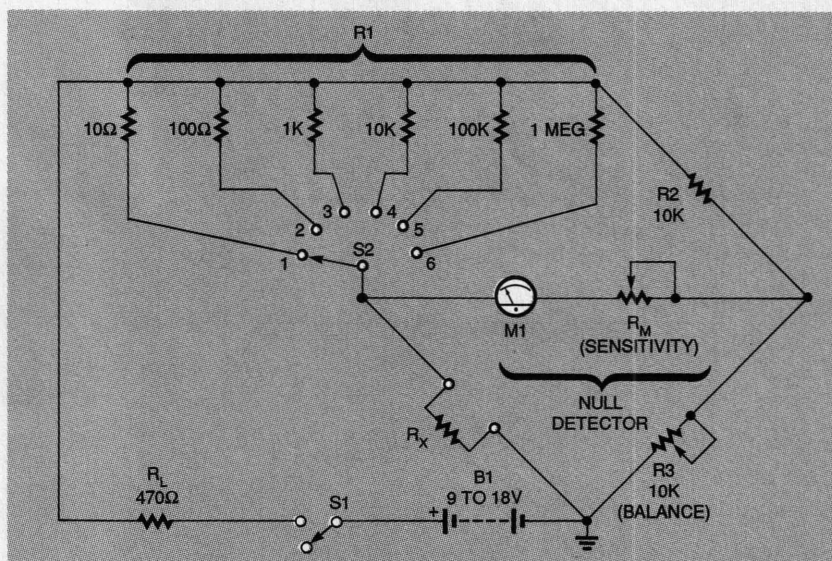


FIG. 6—HIGH-SENSITIVITY, SIX RANGE Wheatstone DC resistance-measuring bridge.

The LF351 has a JFET input for low input offset voltage and BIFET technology provides wide bandwidth and fast slew rates with low bias currents as well as low input offset currents and supply currents. It is the functional equivalent of the LF353, and the LF347 is a dual version.

An external volt-ohmmeter can be set to its 2.5-volt DC range for low-sensitivity measurements, or to its 50-micro-

range version of this circuit, and Table 2 lists its benefits over the circuit in Fig. 3. Its null sensitivity (which is proportional to the $R3/R2$ ratio at balance) is very high on all ranges, and varies from 0.003% at $R3$'s full scale balance value, to 0.03% at one-tenth of full scale.

By confining all measurements to the top nine-tenths of the $R3$ range, one can measure all resistance values in the 1-

TABLE 2
RESPONSE OF 6-RANGE HIGH-SENSITIVITY BRIDGE

Switch S1 Range	Bridge Range	Resistor R1 Value (Ohms)	R1/R2 Ratio (Ohms)	Bridge Null Sensitivity (Nominal)
1	0-10	10	1/1000	Proportional to value of R3
2	0-100	100	1/100	to 0.003% at full scale on all ranges
3	0-1k	1k	1/10	
4	0-10k	10k	1/1	
5	0-100k	100k	10/1	
6	0-1MEG	1MEG	100/1	

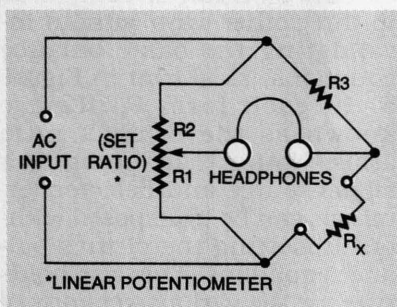


FIG. 7—AC POWERED Wheatstone bridge with variable ratio-arm balancing and headphone-detection.

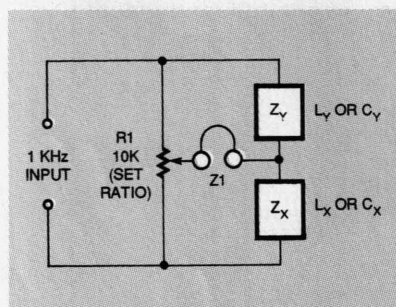


FIG. 9—WHEATSTONE BRIDGE for determining both capacitive and inductive reactances.

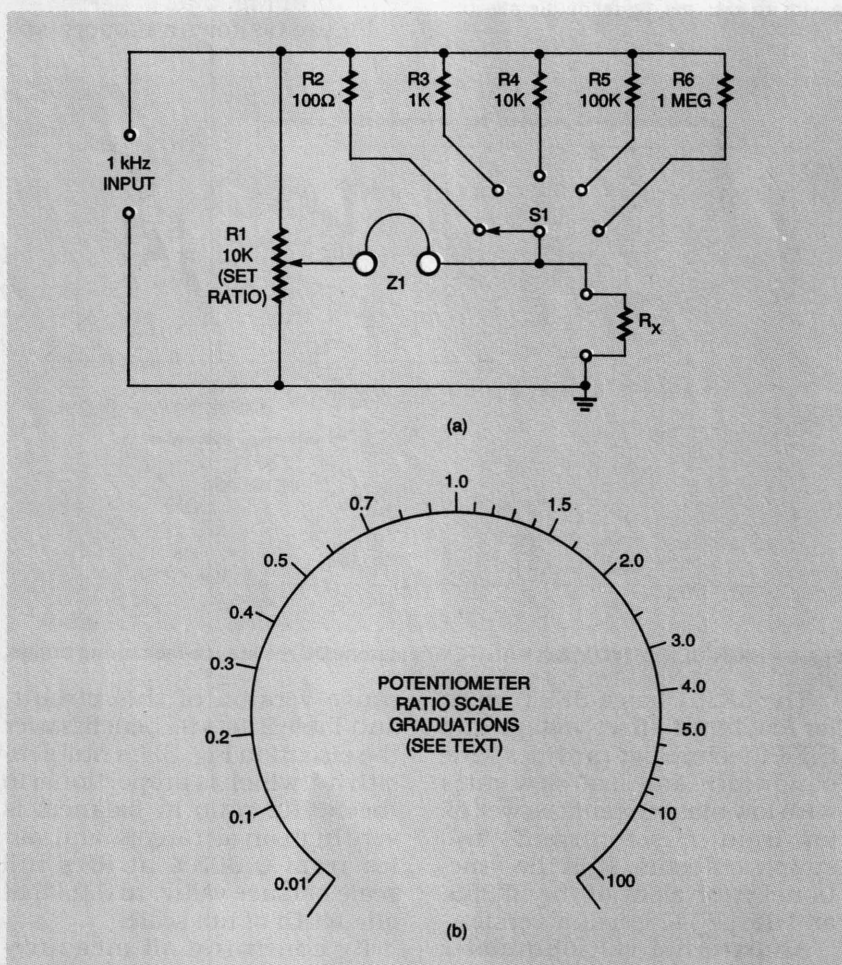


FIG. 8—FIVE-RANGE resistance bridge, with rotary panel potentiometer scale markings (a) and typical ratio scale markings (b).

ohm to 1-megohm range with excellent null sensitivity.

A Wheatstone bridge can be energized from either an AC or DC source without altering the fundamentals of its operation. Figure 7 is the circuit for an AC-energized Wheatstone bridge whose balance condition is maintained with an infinitely-variable pair of "ratio" arms consisting of potentiometer R1 which provides the resistive values of R1 and R2.

The sum of the resistive values of R1 and R2 equals the resistive value of potentiometer. The balance sensitivity of this circuit is high enough to permit detection with headphones.

Figure 8 shows a five-range version of the Wheatstone bridge circuit of Fig. 7. It covers a resistive range of near-zero to near-infinite ohms with its highest precision between 10 ohms and 10 megohms. Rotary panel potentiometer R1's ratio equals one when its wiper is at midrange. The diagram shows expected typical scale graduations for a calibrating this control potentiometer. These must be manually marked as described later in this article.

To measure with the Fig. 8 circuit, connect the bridge to a 1-kHz sinewave generator, insert the unknown resistor R_x , and adjust rotary switch S1 and rotary panel potentiometer R1 until a null can be detected by listening with the headphones. When that occurs, the value of resistor R_x equals the value of the resistor connected to S1 multiplied by the scale value on potentiometer R1. A balance can be obtained on any range, but for the highest precision, the balance should occur with an R1 scale reading between about 0.27 and 3.0.

To calibrate the scale of panel potentiometer R1, insert a 10-kilohm, 1% resistor in the R_x position. Then index rotary switch S1 progressively through its 100-ohm, 1-kilohm, 10-kilohm, 100-kilohm, and 1-megohm positions. Mark the scale at each sequential balance point as 0.01, 0.1, 1.0 (mid-scale), 10, and 100.

Repeat this procedure with

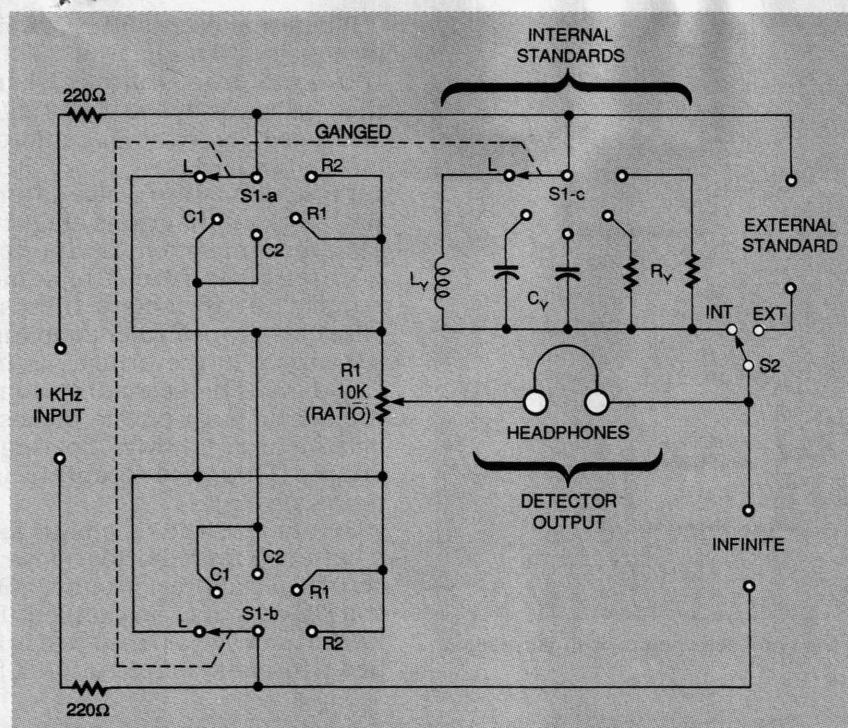


FIG. 10—LCR BRIDGE with headphone null detector.

R_X values that are multiples or submultiples of 1.5, 2, 3, 4, 5, etc., until the scale is completely calibrated, as shown in Fig. 8-b.

Resolution and precision

There are four important quality characteristics of a measurement bridge: (1) measurement range, (2) balance-sensitivity, (3) resolution, and (4) precision. The term resolution refers to the accuracy with which the R_X value can be read from the bridge's controls. For example, in Figs. 3 and 6, R3 gives a resolution of about 1% of full-scale if it is a manually-calibrated, linear, potentiometer, or about 0.005% of full-scale if it includes a four-decade resistor box.

The resolution of the bridge in Fig. 8 varies from $\pm 1\%$ at a "1" ratio to $\pm 2\%$ at a 0.3 or 3.0 ratio, to $\pm 5\%$ at a 0.1 or 1.0 ratio.

The term precision refers to the basic accuracy of the bridge, assuming that it has perfect balance-sensitivity and resolution, and equals the sum of the R1/R2 ratio tolerance and the tolerance of the resistance standard R3. If the R1/R2 ratio is set strictly by precision resistors, the ratio's precision equals the

TABLE 3
STANDARDS VS. BRIDGE VALUE

Standards Value	Bridge Range
R3	100 Ω 10 Ω -1 k
	100 k 1 k-100 k
	1 MEG 100 k-10 MEG
C3	100 pF 10 pF-0.001 μ F
	0.01 μ F 0.001 μ F-0.1 μ F
	1 μ F 0.1 μ F-10 μ F
L3	1 mH 100 μ H-10 mH
	100 mH 10 mH-1 H
	10 H 1 H-100 H

sum of the R1 and R2 tolerances. However, there are other techniques that permit resistors to be matched so that ratio errors are reduced to only about $\pm 0.005\%$.

For example, if the high-resolution Wheatstone bridge in Fig. 6 is made with R1 and R2 as 1% resistors and R3 is a manually-calibrated control potentiometer, the circuit will have a basic precision of only 3%. However, if the values of R1 and R2 are correctly matched, the precision of the bridge increases to $\pm 1.005\%$.

The circuit's precision can be increased to $\pm 0.105\%$ with a $\pm 0.1\%$ multidecade resistor box in the R3 position. Bear in

mind that additional errors can creep in when you are measuring very low or very high resistance values. The most likely cause will be the resistance values of switch contacts and leads when measuring low resistive values, and leakages when measuring high values.

The overall quality of a bridge depends on its balance-sensitivity, resolution, and precision. Thus, the circuit in Fig. 6 offers excellent sensitivity and very good resolution and precision. Consequently, it can function as the basic circuit of either an inexpensive, simple bench test instrument or as a component in a precision laboratory instrument, depending on how it is built.

By contrast, the bridge in Fig. 8 has intrinsically poor resolution and precision. This means that it is suitable only for fast, imprecise measurements such as might be satisfactory for servicing equipment.

Capacitance and inductance

An AC-energized Wheatstone bridge can measure reactance, resistance, capacitance (C) and inductance (L). The circuit in Fig. 9 is a modification of the circuit in Fig. 7 to measure C or L values by replacing R4 and R_X with comparable reactances. This will work if C_X or L_X are reasonably pure and have 1-kHz impedances greater than about 1 ohm and less than 10 megohms.

The difficulty in trying to measure inductance with this circuit is that accurate inductors (for use in the Z4 position) are hard to find, and inductive impedances are only 6.28 ohms per millihenry at 1 kHz.

The only difficulty in measuring capacitance is that the C_X value is proportional to the reciprocal of potentiometer R1 resistance scale markings. If the basic bridge is to measure both R and C, two calibrated sets of potentiometer R1 scales are needed. This drawback can be overcome by fitting R1 with a reversing switch. An example of this is shown as the multi-range LCR bridge of Fig. 10. Thus, only a single scale as

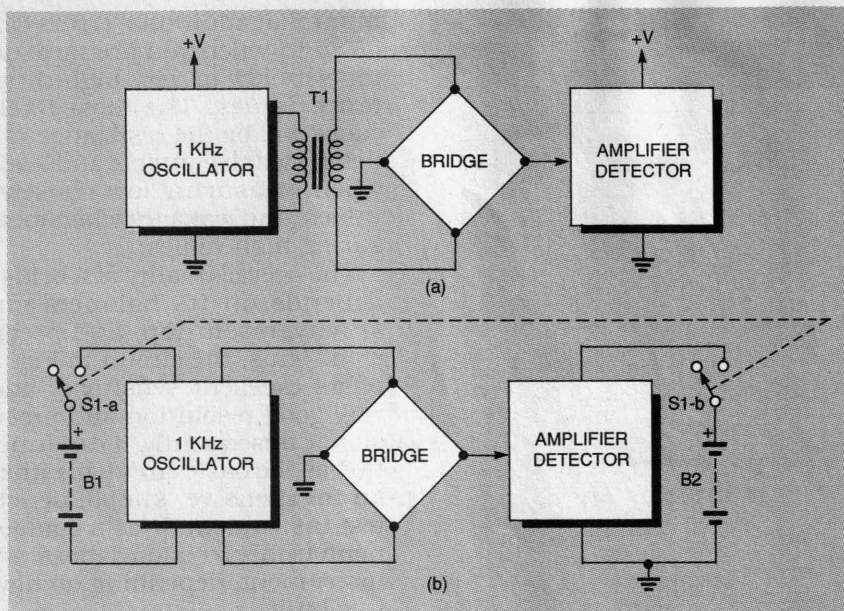


FIG. 11—ALTERNATIVES CIRCUITS for powering and detecting nulls in Wheatstone bridge circuits.

When that occurs, the C_x value equals 100 nanofarads. This 100-nanofarad standard can then be connected to the bridge and used to create a 1 microfarad standard.

Figure 11 shows two alternative block diagrams. Figure 11-a, the first option, is to power both circuits from the same supply but to isolate the oscillator by transformer-coupling its output to the bridge, as in Fig. 11-a. The second option, Fig. 11-b, is to power the oscillator from its own "floating" supply. This second option is more efficient.

Figure 12 is the schematic for a battery-powered bridge power source. It can provide either a 9-volt DC output or an excellent 1-kHz sinewave output with a peak-to-peak amplitude of 5

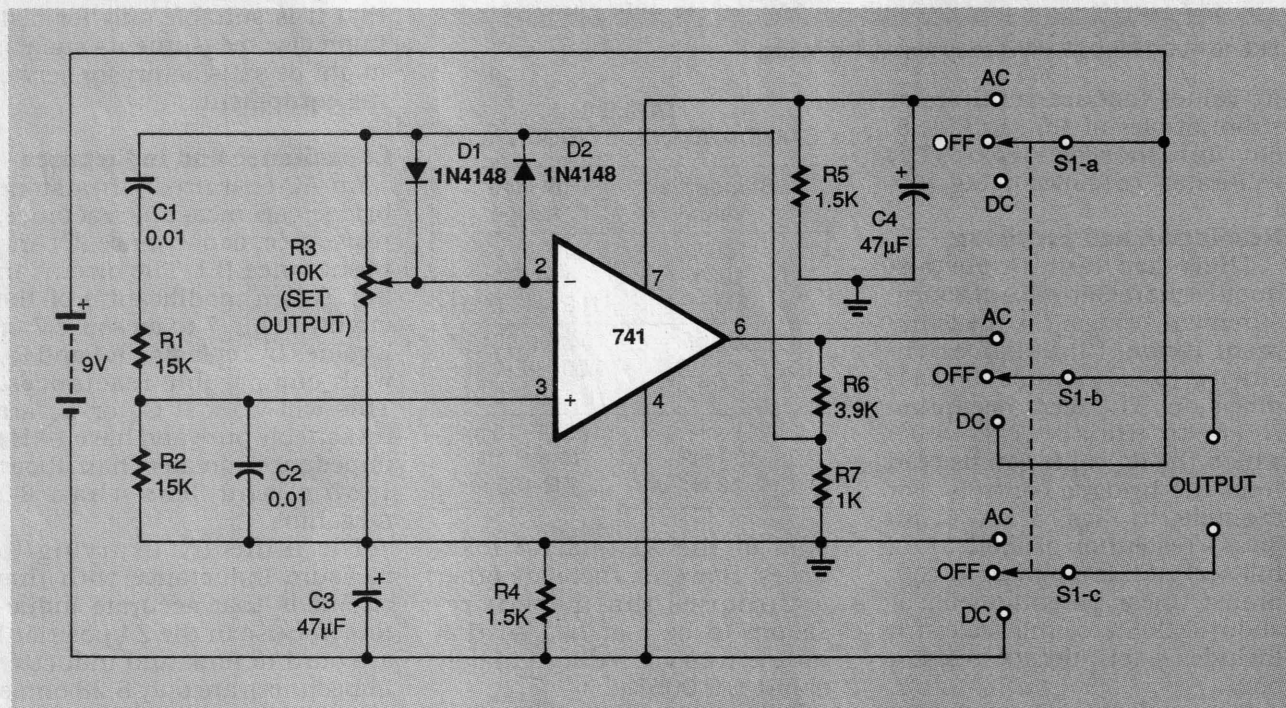


FIG. 12—POWER SOURCE for bridge circuits capable of 9-volt DC and 5-volt, 1-kHz outputs.

shown in Fig. 8 is required.

The general purpose LCR bridge with headphone detector shown in Fig. 10 is a versatile instrument. Switch S2 permits it to be used with either internal or external L, C, or R standards. The midscale value of each range is equal to the value of the standard assigned for that range, as listed in Table 3.

When this instrument is cali-

brated, it can be help to create its own alternative measurement standards. For example, if an accurate 10-nanofarad capacitance standard is inserted across the EXTERNAL STANDARD terminals, a 100-nanofarad standard can be created by moving potentiometer R1 to the "10" position and then connecting capacitors in parallel across the "X" terminal for a null balance.

volts. The oscillator is a diode-stabilized Wien bridge oscillator that works from a split power supply derived from the battery through resistors R1 and R2.

The circuit has a low-impedance output, and consumes a less than 4 milliamperes of quiescent current. To set up the oscillator, connect its output to a suitable oscilloscope and trim potentiometer R1 to give a reasonably pure sinewave output of about 5 volts, peak-to-peak. Ω



Audio Gets Easier

Using
the
New
LM4862

MICHAEL A. COVINGTON

Any electronic device that makes sound in a speaker—at least if the sound is more than digital beeps—needs an audio amplifier. Until now, the most popular IC for driving small speakers has been the LM386, introduced in 1975 by National Semiconductor. Figure 1 shows a typical circuit. Compared to the discrete-component circuits that preceded it, the LM386 was a godsend; it could deliver 0.2 watt into an 8-ohm speaker from a 9-volt supply without an output transformer.

But the LM386 wasn't perfect. As Fig. 1 shows, it requires large electrolytic capacitors, which add bulk and cost to the circuit and can distort the sound as they age. Further, the input impedance of the LM386 is

This new "Boomer" audio amp from National Semiconductor is easier to use, offers better performance, and needs fewer bulky external components than anything previously available.

very high, making it prone to oscillation if the inputs aren't carefully separated from the outputs. Its voltage gain of 20 (or 200 with one capacitor added) is a bit too high for line-level input (1V RMS) and contributes to the oscillation problem.

Twenty years went by, and National Semiconductor has recently introduced a new line of easy-to-use "Boomer" audio amplifiers. The handi-

est of these is the LM4862, which is available in both surface-mount and DIP packages with the same pinout (LM4862M and LM4862N respectively). It can deliver 0.675 watt into an 8-ohm speaker at 1% total harmonic distortion. (At slightly lower power levels, distortion is appreciably less.) What's more, it has automatic thermal shutdown to protect it from damage if overloaded, and it operates from a single 5-volt supply. The input impedance is relatively low and user settable, so the risk of oscillation is much lower.

Inside the Chip. Figure 2 shows what's inside the device. The LM4862 drives the speaker differentially, applying opposite waveforms to the two terminals—a configura-

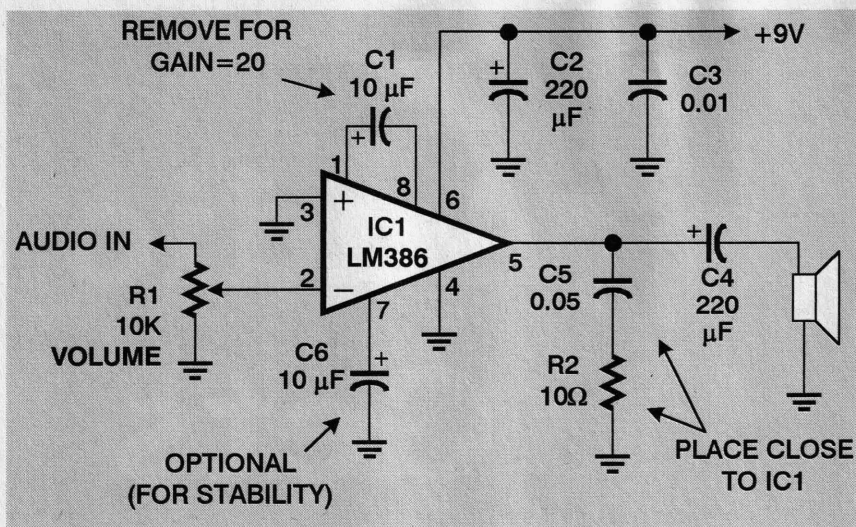


Fig. 1. The classic LM386 audio amplifier needs six capacitors, three of them electrolytic.

tion often called BTL (bridge-tied load). Either side of the speaker can go to +5V, while the other side

goes to 0V. Thus, the amplifier can produce 10 volts of swing from a 5-volt supply. That's enough for room-

filling volume with a 4-inch full-range speaker.

The supply voltage can range from 2.7 to 5.5 volts; naturally, the amplifier can deliver more power if the supply voltage is near the high end of the range. You can power the LM4862 from two or three 1.5-volt cells or a regulated 5-volt supply. (Do not use a 6-volt battery; it exceeds the maximum rating.) Total current consumption ranges from 5 mA when quiet to about 250 mA at maximum volume. Power-supply ripple rejection is excellent, better than 50 dB when $C2 = 1 \mu\text{F}$.

Using the Chip. A typical amplifier circuit is shown in Fig. 3. It's simple and does not require electrolytic capacitors; that keeps the cost down and ensures high-fidelity audio. Actually, $C2$, the bias bypass

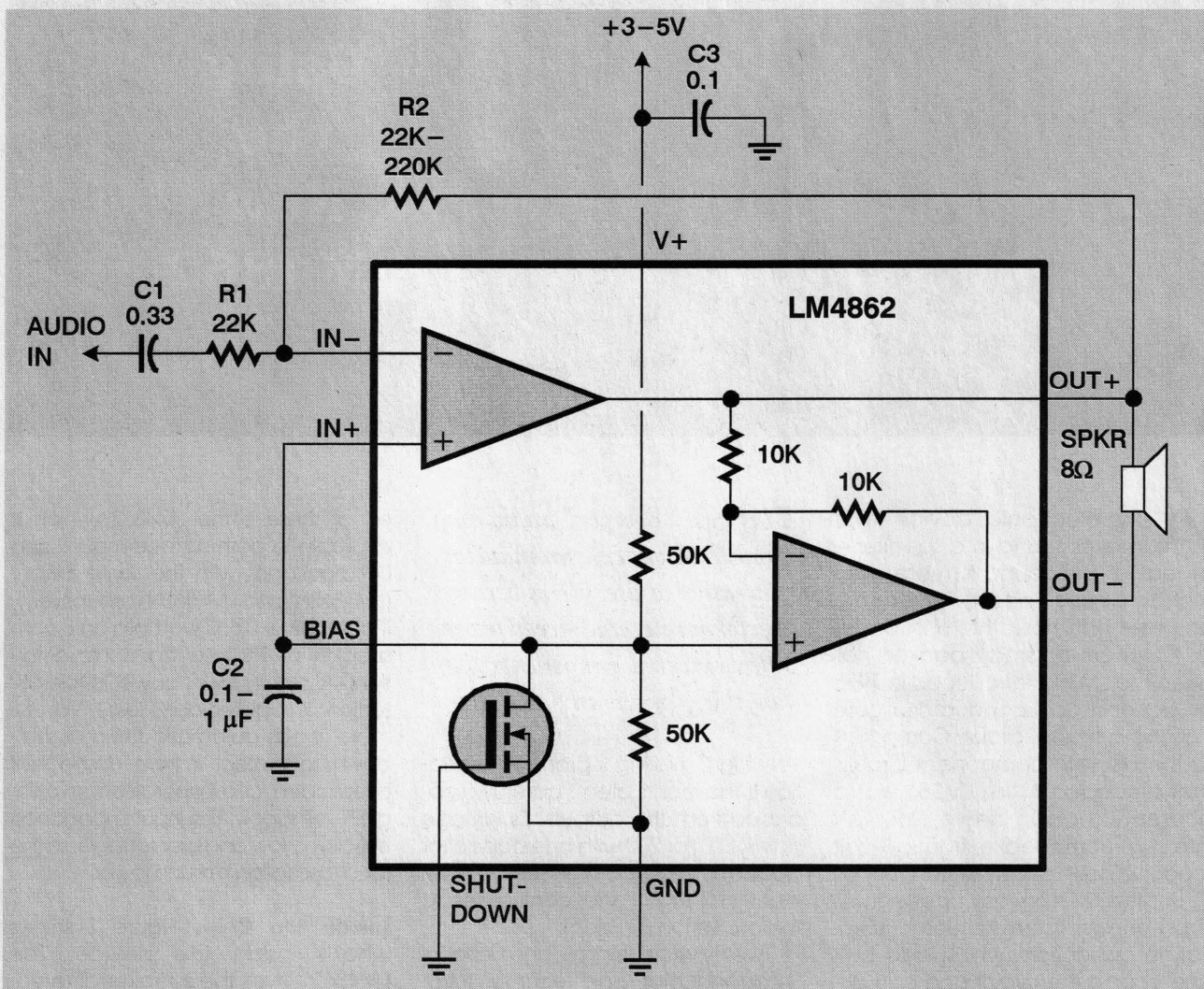


Fig. 2. The new LM4862 needs fewer external components and is less prone to oscillation than the LM386.

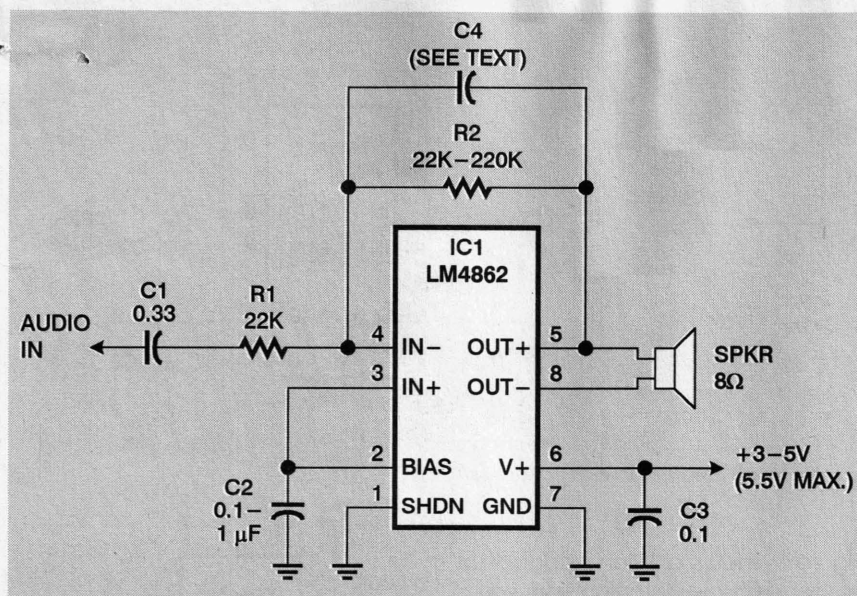


Fig. 3. A typical circuit using the LM4862. Place C3 as close to the IC as possible.

capacitor, can be a tantalum electrolytic; the audio signal doesn't pass through it. Also, a 100-μF electrolytic in parallel with C3 is a good idea when using batteries or a poorly regulated power supply.

The voltage gain is equal to $2(R2/R1)$ and should not exceed

	HIGH FIDELITY (20-20,000 Hz)	POWER SAVING (150-20,000 Hz)
LOW GAIN ($A_v = 2$)	R1 = 22K R2 = 22K C1 = 0.33 C2 = 1.0 μF C4 OMITTED	R1 = 22K R2 = 22K C1 = 0.05 C2 = 0.1 μF C4 OMITTED
HIGH GAIN ($A_v = 20$)	R1 = 22K R2 = 220K C1 = 0.33 C2 = 1.0 μF C4 = 22 pF	R1 = 22K R2 = 220K C1 = 0.05 C2 = 1.0 μF C4 = 22 pF

Fig. 4. Here are some component values for variations of the circuit in Fig. 3.

20; sound quality is best when $R2 = R1$ and the gain is 2. (That's exactly what you need to drive a speaker from 1-volt line-level or headphone-level audio.) When the gain is higher than about 5, $R2$ should be bypassed by a small capacitor ($C4$) to prevent oscillation. A 5-pF capacitor will do the job, but a 22-pF capacitor will probably be easier to find; do not use values larger than 32 pF. If the input signal is coming from a low-impedance source, you can use smaller resistors, such

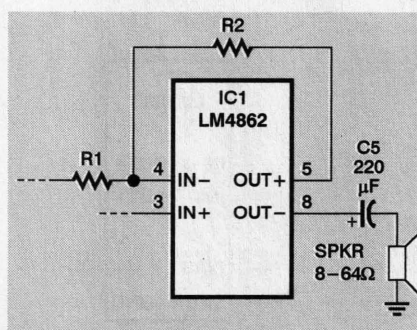


Fig. 5. How to drive a grounded circuit. The balance of the circuit is shown in Fig. 3.

as $R1 = 4.7K$ and $R2 = 4.7K$ to 47K.

Figure 4 gives component values for some typical amplifiers. Note in particular that you can save battery power, as well as component cost, by cutting the bass response when driving a small speaker that wouldn't reproduce low frequencies anyhow.

The speaker impedance must be at least 8 ohms. You can use a 16-, 32-, or 64-ohm speaker, but you'll get considerably less power out. If you must ground one side of the speaker, drive the other side through a capacitor as shown in Fig. 5; you'll get only a quarter as much power as if the speaker were driven differentially.

The shutdown (SHDN) pin disables the amplifier, bringing both outputs low and cutting power consumption to less than 1 micro-ampere (yes, microampere), when connected to +5V. The shutdown pin is grounded in normal use. You can use it to implement a "mute" button without putting a switch in the signal path, or even wire it to a headphone jack to silence the speaker when headphones are plugged in.

The bias pin is the output of a voltage divider built into the chip. Its purpose is to hold the positive inputs of both op-amps at half the supply voltage so that they can operate with a single supply. You can also use the bias pin to bias one or two additional op-amps as shown in Fig. 6.

The bias pin should be bypassed to ground by a capacitor between 0.1 and 1.0 μF. Larger values improve ripple rejection and suppress the "thump" when the amplifier is turned on.

Some Applications. The LM4862 teams up well with other low-voltage ICs. For example, Fig. 7 shows an experimental AM radio based on the Plessey (formerly Ferranti) ZN414 TRF receiver chip. On the breadboard, this circuit gave high-fidelity

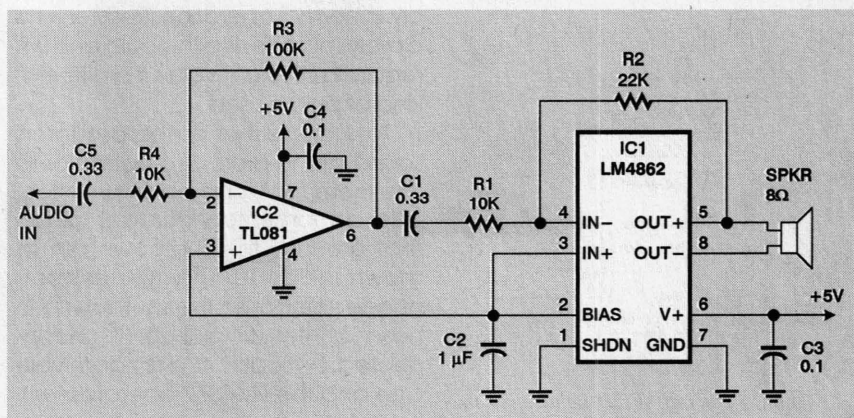


Fig. 6. You can use the bias pin, pin 2, to bias one or two external op-amps as shown here.

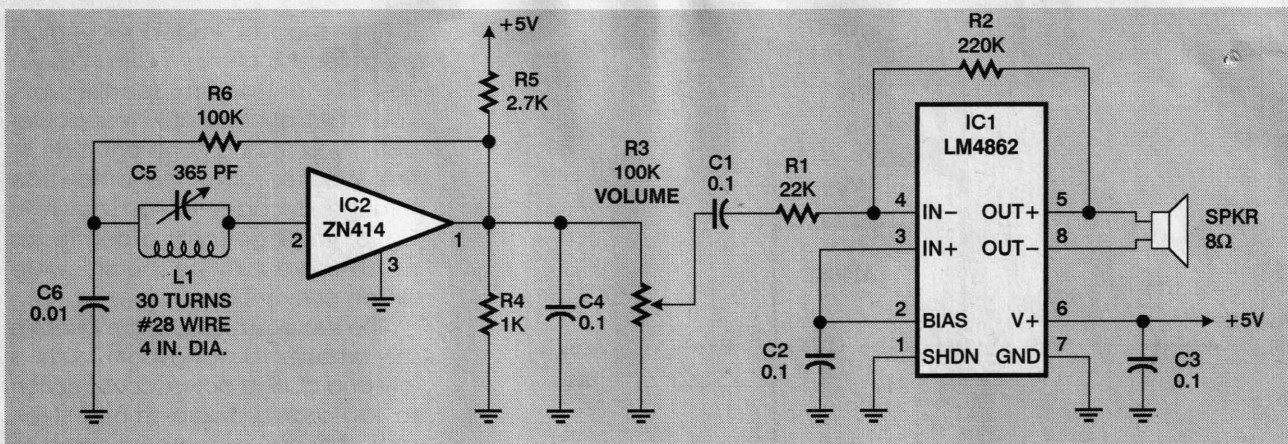


Fig. 7. When the LM4862 is combined with the Plessey (Ferranti) ZN414 receiver chip, the result is a simple, high-quality AM receiver.

reception of local AM stations.

Like other amplifiers, the LM4862 makes a fine oscillator. Figure 8 shows a classic op-amp square-

Figure 9 shows a Wien bridge that produces a low-distortion sine wave.

In the last circuit, a 1.5-volt, 15-mA light bulb (RadioShack 272-1139)

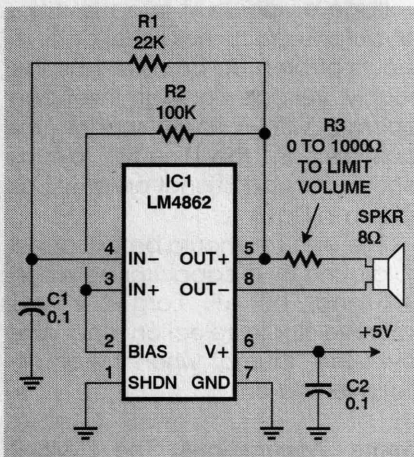


Fig. 8. This squarewave oscillator produces a loud 1-kHz tone in the speaker.

wave oscillator that can serve as a loud siren—at least twice as loud as any conventional 5-volt circuit.

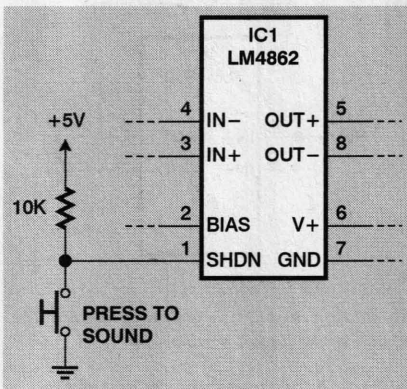


Fig. 10. The oscillators in Figs. 8 and 9 can be pushbutton controlled as shown here.

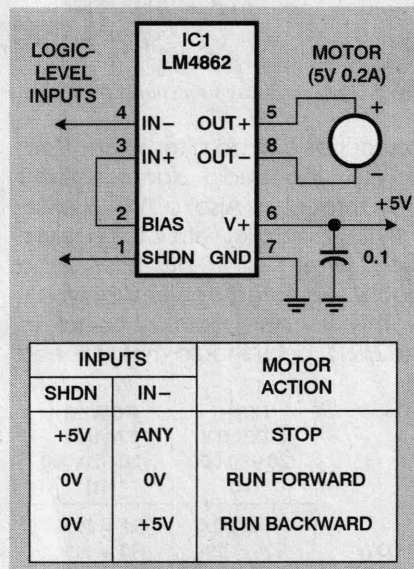


Fig. 11. Using the LM4862 as a bi-directional motor controller.

serves as the regulating element in the feedback loop. As current increases, the bulb heats up and its resistance increases, cutting feedback and stabilizing the oscillator. The bulb does not actually glow visibly in normal operation. Resistor R1 is closely matched to the bulb and will require a different value if a different kind of bulb is used.

To make either of the oscillators sound at the press of a button or to use them for Morse code practice, add a momentary-contact switch that grounds the shutdown pin as shown in Fig. 10. The 10K resistor is needed because the shutdown pin has no internal pull-up; if unconnected, it "floats" to a random voltage and the LM4862 operates very erratically.

Finally, a speaker isn't the only

thing the LM4862 can drive differentially. It also makes a fine full-bridge bi-directional driver for small motors. Figure 11 shows a circuit that was breadboard-tested with a 6-volt tape-recorder motor (running on 5 volts, of course). In this circuit, the inputs of the LM4862 are compatible with CMOS logic outputs, making computer control easy to implement.

Availability. One potential problem with the chip for hobbyists is that it is not that widely available. National does provide free samples to engineers and sometimes others. For more information, see their Web site at www.national.com. It is also available from traditional full-line distributors such as Newark Electronics (www.newark.com). Ω

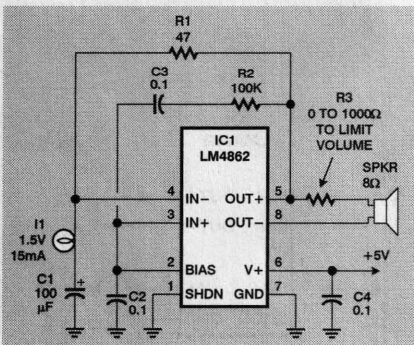


Fig. 9. A tiny "grain-of-wheat" light bulb is the feedback regulator in this circuit for a Wien-bridge oscillator.

Listen to the Earth with this Seismic Detector



RON NEWTON

Want to find out what's happening underfoot? You can do just that with this inexpensive unit.

If you live in a seismically active area, you are no doubt aware of the "non-permanence" of the ground that you are standing on. Did you ever wonder if you have just felt a small earthquake, or was it only a passing truck? Perhaps you would like to explore the concept of volcanic eruption predictions.

The human body can detect a seismic event of about 2.5 or above on the Richter scale. A seismograph would be nice, but who can afford to have reams and reams of paper spewing out on the floor day after day? Besides, commercial seismographs are very expensive instruments and most are not portable.

To the rescue comes the Seismic Detector presented here, a portable and paperless device that is inexpensive and versatile. In addition to seismometers, the basic circuit can be used with other transducers such as LVDTs, differential pressure sensors, and accelerometers to name a few. Even with the use of expensive low-power components,

the cost for parts is currently about \$40; lower-cost standard components can be used if battery life is not a concern.

Design Features. Several features of the Seismic Detector contribute not only to its low cost but its usefulness as a geologist's instrument. Sensitivity is less than 2 on the Richter scale. At that level, it will be able to detect and record seismic events that we wouldn't notice. Up to eight separate one-minute events can be recorded and stored by the unit. Once an event has been detected and recorded, a light-emitting diode indicates that a seismic event has taken place and is stored in the system's memory.

The Seismic Detector can be

connected to a PC for real-time monitoring of seismic events. That does not mean, however, that the Seismic Detector is restricted to any area that has a convenient wall outlet. Designed for portability and use in the field, it can run for about one year on a fresh set of batteries. Since it can collect data that might occur months apart, the date and time of each event is also stored by the unit.

Naturally, the computer interface can also be used for downloading any stored information for plotting and analysis.

Circuit Description. The schematic diagram of the Seismic Detector is shown in Fig. 1; follow it during the following discussion.

The heart of the system is SEN1, a vibration transducer. Manufactured by Geophone for use by the oil industry, it is a small can measuring 1 inch in diameter by 1.3 inches in height. It has a natural frequency of 10 Hz. If you open up a sensor, you'll find that it consists of little more

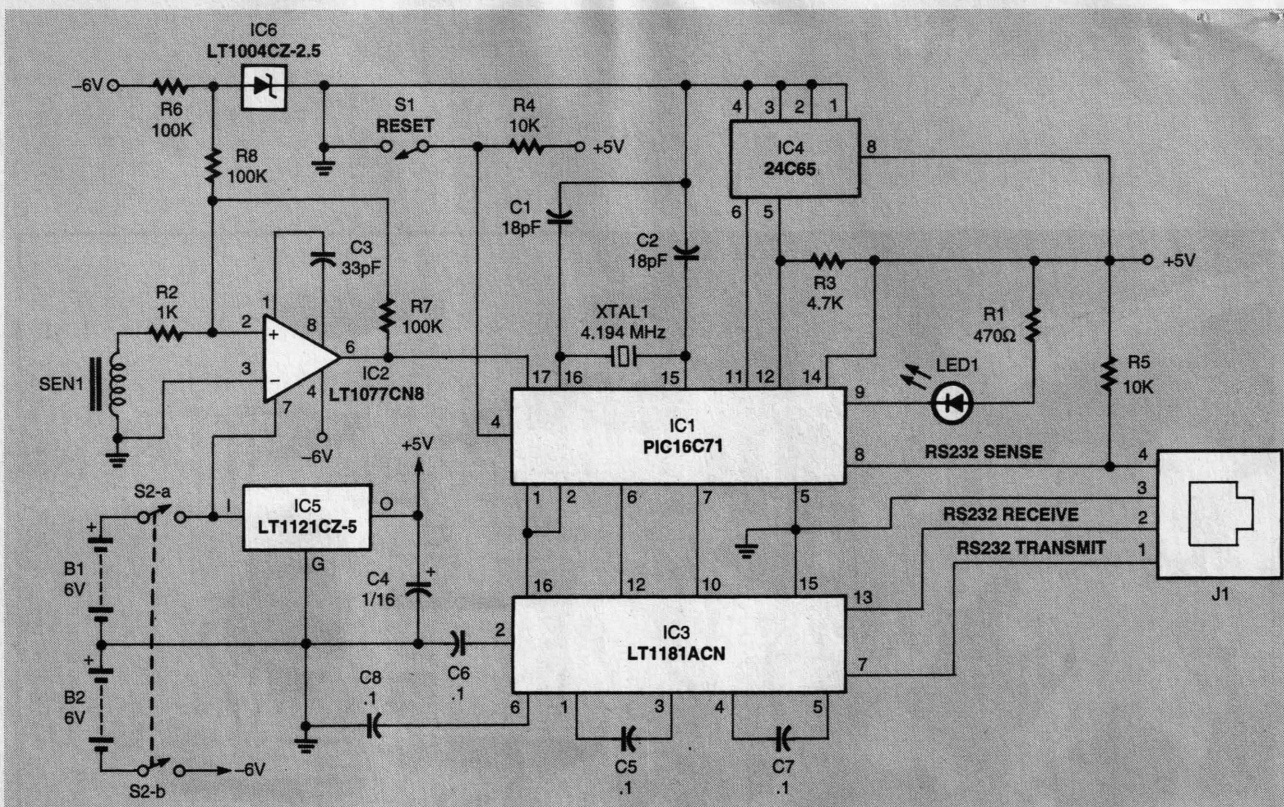


Fig. 1. Using low-power components to extend its battery life to a year of continuous operation, the Seismic Detector can store eight separate seismic events. Enough memory is provided for a minute's worth of samples per event—at 16 samples per second.

than a coil with a magnet suspended by a spring. When the sensor shakes, the magnet moves up and down within the coil. As everyone remembers from basic electricity, an electric current is generated when a conductor passes through a magnetic field. The sensor uses that effect to generate an AC volt-

age when it is moved. The intensity and shape of the AC signal is related directly to the amount of vibrational shock detected.

The output from SEN1 is very weak; IC2 amplifies it 100 times to boost the Seismic Detector's sensitivity. The amplified signal is applied to one of the inputs of IC1, a

PIC16C71. That particular microcontroller features an on-board 8-bit analog-to-digital converter. There is a limitation on the input signal to the A/D converter—it cannot go negative, making measuring an AC waveform a bit difficult. The solution is to raise the voltage level of the signal so that it doesn't exceed the input specifications of the 16C71. Take another look at IC2; it is set up as a summing amplifier. A 2.5-volt offset is added to the sensor signal by using a precision reference diode, IC6. That voltage is at the halfway point on the A/D converters limits (128 on a 0-255 scale). Thus, a negative voltage will read

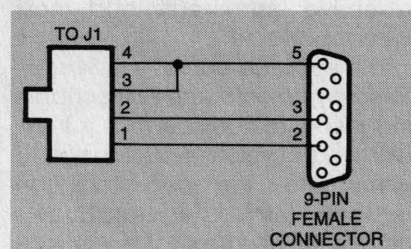
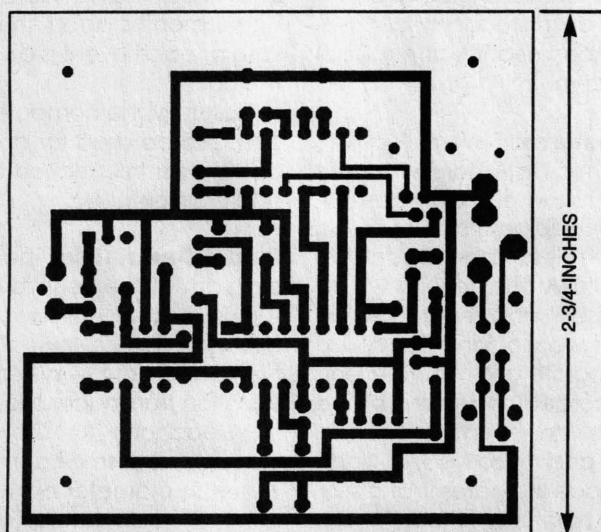


Fig. 2. This cable connects the Seismic Detector to a PC for downloading the recorded data as well as setting up the unit.

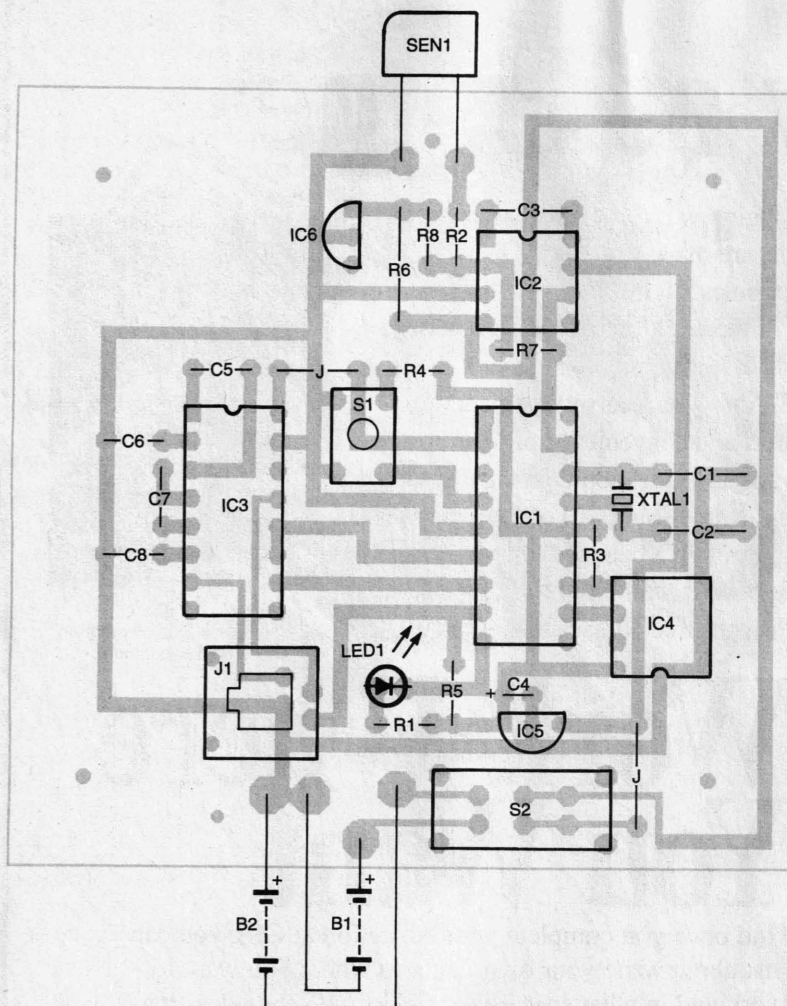


Fig. 3. The Seismic Detector is a simple circuit that fits on a small single-sided board. Only two jumpers are needed.

between 0 and 127; positive voltages will be between 129 and 255.

The frequency of XTAL1 was chosen so that together with the software that is programmed into IC1, precise timing lengths can be generated. For example, IC1 samples its analog/digital converter 16 times per second. If the reading is less than 123 or over 133, the software turns on LED1, which will flash every 15 seconds. One minute's worth of readings are stored in IC4, a serial memory chip. After a total of eight events, the system is set to "go to sleep" without taking any more readings no matter what happens. Due to the nature of how IC4 stores information, no data will be lost if power is lost. However, the unit will not take any more readings even if power is restored.

A bi-polar power supply is needed because of IC2. The positive side

is supplied by four D-size batteries and is regulated to 5 volts by IC5, a low-power voltage regulator. A set of four AA cells provides the negative voltage for IC2.

Communications to a PC is handled by IC3, an RS232 interface chip. To conserve power, IC1 only activates IC3 when an adapter cable is plugged into J1. That adapter cable, shown in Fig. 2, shorts pins 3 and 4 of J1 together; IC1 watches for that condition. The Seismic Detector uses a 9600-baud rate to keep it compatible with the oldest of PCs.

Construction. The Seismic Detector's circuit is simple enough to build on a piece of perfboard using standard construction techniques. However, using a PC board makes for a neater appearance and eliminates most wiring errors.

If you choose to use a PC board, a foil pattern for a single-sided board has been included here. Alternatively, an etched and drilled board is available from the source given in the Parts List.

Before beginning assembly, note that IC1 must be programmed with the software that runs the Seismic Detector. The software for the Seismic Detector can be downloaded from the Gernsback FTP site (<ftp.gernsback.com/pub/EN/seismic.zip>). If you do not have access to a PIC programmer, a pre-programmed chip can be purchased from the source given in the Parts List, but download the software anyway. That package includes the PC software for reading back the information that the Seismic Detector has collected.

Start construction by placing the D-cell battery holder upside down with the wires to the left. Mount the AA-cell battery on the right side of the D-cell holder using double-sided tape or other suitable adhesive; the wires for both battery holders should face the same way.

Place the PC board on the left side of the D-cell battery holder. Mark the four locations where screws will hold the PC board to the holder. Drill those holes in the holder and set it aside.

If you are using a PC board that has been purchased from the source given in the Parts List or have etched one from the foil pat-

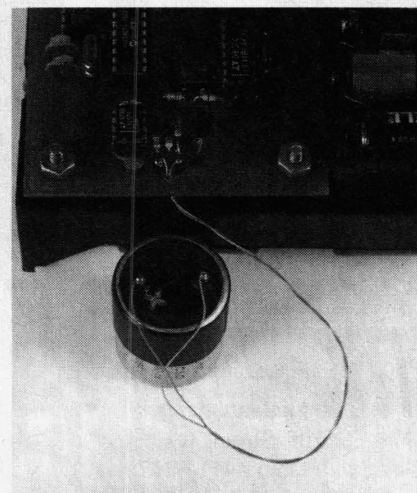


Fig. 4. The Seismic Detector relies on a vibration sensor that is used by the oil industry for exploration.

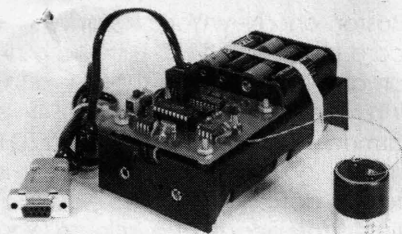


Fig. 5. The completed Seismic Detector is a compact, self-contained unit.

tern, use the parts-placement diagram shown in Fig. 3 to locate the various components. Note that there are two jumpers; they can be made from scrap pieces of component lead. In general, start with the smallest components and work your way up to the larger devices. Before soldering any polarized components, be sure to double-check their orientation. Burned semiconductors let you know that they have been installed backwards by releasing a puff of smoke and a unique odor that is unlike any other smell on Earth. Electrolytic capacitors, when installed the wrong way, have a way of failing that is—in a single word—“thrilling!”

A pair of square wire-wrap posts should be installed where SEN1 will be connected to the board. The integrated circuits can be soldered directly to the board; sockets can be used, but they might cause reliability problems after the Seismic Detector has been out in the field for several months exposed to extremes of temperature and humidity. Note the orientation of IC4; it is the reverse of the other chips.

You might want to consider a socket for IC1, especially if you are using the re-programmable version and might replace the program with one of your own design. Again, reliability and upgradability are the two features that you must choose between when it comes to the use of IC sockets.

When the board is done, look over your work carefully for poor soldering or missing, backwards, or misplaced components.

Note that there is a large strain-relief hole next to J1. Run the battery wires through that hole and solder them to the appropriate pads. The red wire from the AA batteries goes to ground and its black

wire to the negative-voltage input. On the D-cell wires, the red goes to the positive-voltage input and the black to ground.

The transducer is connected to the board in a similar way; a strain-relief hole is next to the pads for the transducer wires. That hole is much smaller—use two different colors of wire-wrap wire. Twist the leads together and keep their length less than 12 inches to prevent any electromagnetic interference (EMI). The wiring method used in the author's prototype is shown in Fig. 4.

Mount the board to the back of the battery holder using standoffs, screws, and nuts. That completes assembly of the Seismic Detector itself.

Cut one end off of a modular 4-conductor telephone handset cord. Strip and tin the wires on the cut end. Following the schematic diagram shown in Fig. 3, connect the wires to a 9-pin connector. Use an ohmmeter or continuity tester if you are unsure as to which wire goes to which connector pin. The transmit line goes to pin 2 and the receive line to pin 3. Twist the ground and RS232 sensor wires together and solder them to pin 5.

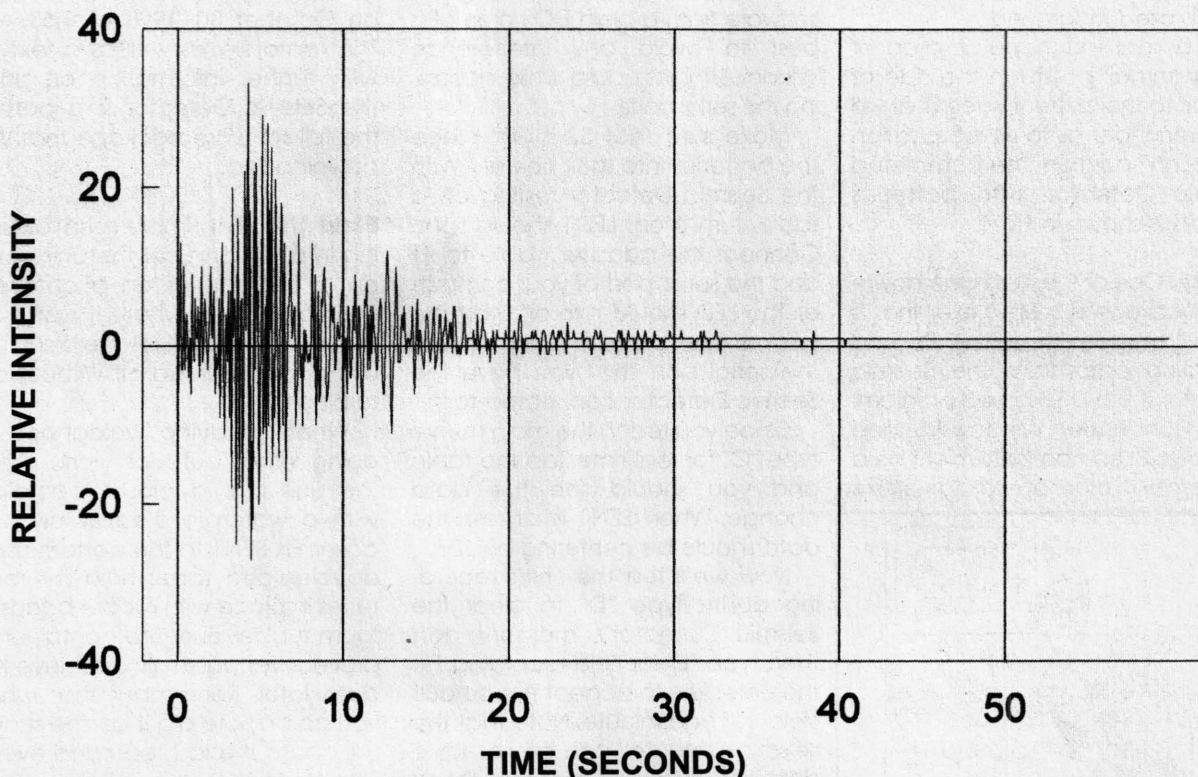


Fig. 6. An example of recorded data from the Seismic Detector is this plot of an earthquake that occurred in Carson City, NV on October 30, 1998.

PARTS LIST FOR THE SEISMIC DETECTOR

SEMICONDUCTORS

IC1—PIC16C71 microcontroller, integrated circuit
 IC2—LT1077CN8 op-amp, integrated circuit
 IC3—LT1181ACN RS-232 driver, integrated circuit
 IC4—24C65 serial EEPROM, integrated circuit
 IC5—LT1121CZ-5 low-power 5-volt regulator, integrated circuit
 IC6—LT1004CZ-2.5 micropower 2.5-volt reference, integrated circuit
 LED1—Light-emitting diode, red

RESISTORS

(All resistors are 1/8-watt, 5% units.)
 R1—470-ohm
 R2—1000-ohm
 R3—4700-ohm
 R4, R5—10,000-ohm
 R6—R8—100,000-ohm

CAPACITORS

C1, C2—18-pF, ceramic-disc
 C3—33-pF, ceramic-disc
 C4—1-μF, 16-WVDC, electrolytic
 C5—C8—0.1-μF, polyester

ADDITIONAL PARTS AND MATERIALS

B1, B2—6-volt battery
 J1—modular telephone connector, PC-mount
 SEN1—Seismic vibration transducer (see text)
 XTAL1—4.194-MHz crystal
 9- or 25-pin connector, four D-cell battery holder, wire hardware, etc.

Note: The following items are available from Tahoe Chemical R&D; Tel: 702-885-8842; E-mail: sjnewt@aol.com: Pre-programmed IC1, \$10; etched and drilled PC board, \$15. Please add \$5 for shipping and handling within US and Canada. NV residents must add appropriate sales tax.

SEN1 is available from: All Electronics, PO Box 567, Van Nuys, CA 91408-0567; Tel: 800-827-5432; item GP-1; and from Geo Space Corporation, 7334 N. Gessner, Houston, TX 77040; Tel: 713-939-7093; Web: www.geospacecorp.com; item GS-20DM.

If you need to use a 25-pin connector, pins 2 and 3 are reversed; pin 7 is used for ground.

As a final touch, put a drop of red fingernail polish on the side of S2 next to the battery wires. That will be a reminder as to which position is the "on" position. The completed Seismic Detector, with batteries loaded, is shown in Fig. 5.

Testing. One of the files in the download package is seismic.exe; that is the PC program for setting up and collecting data from the Seismic Detector. It is a DOS-based program; it will run under Windows95 and Windows98; compatibility might be a problem with other operating systems

such as WindowsNT, Windows2000, and OS/2. At any rate, running the software from a plain DOS prompt is best to avoid any interference Windows might cause while accessing the serial ports.

Make sure that S2 is off; place the batteries into their holders. With the Seismic Detector resting on a table, turn it on; LED1 should light. Connect the adapter cable to J1 and the serial port of your computer. The LED should turn off. Start the seismic.exe program and select the serial port that you have the Seismic Detector connected to.

Once you reach the main menu, type "C" for real time. Tap the table and you should see the data change. When SEN1 is at rest, the data should be centering on 128.

Now we'll test the unit's recording ability. Type "D" to clear the Seismic Detector's memory, and then type "E" for field recording. Hit the enter key to accept the default delay of one minute. Note that the Seismic Detector can be set for a delay of up to 255 minutes—that's four hours and 15 minutes, the approximate driving time between

Boston and New York! Remove the cord from J1 and LED1 should blink on and off indicating that it is arming. Once LED1 goes out, it is armed. Tap the table and LED1 should go on indicating that it is taking data. Tap the table several times for about one minute. You can turn S1 off; the unit will retain its data.

Downloading. Plug the adapter cable into J1 and turn the unit on. At the main menu of the PC program, type "B" to choose the store-to-disk option, then type "A". Collection will automatically stop when all of the available data is downloaded. Once you are satisfied that you have stored the data you want, you can erase the memory as detailed above.

The data files can be read into any spreadsheet. Both Lotus 123 and Excel work well. The first large number is the date and its decimal is the time that the seismic event took place. The following data is seismic data. Use the spreadsheet's graphing function to display the data. An example of graphing a seismic event is shown in Fig. 6. That event took place in Carson City, NV on October 30, 1998 at 1:53 AM. That data file, as well as a text file with further information on using the Seismic Detector, is a part of the full software package that you downloaded.

Field Use. For simple seismic sensing in a building, set the unit down on its D batteries on a concrete floor in an area where it won't be disturbed. Place SEN1 next to it. The unit can be turned off without disturbing SEN1.

When studying volcanoes or doing other outdoor work, place the unit into a plastic container with a waterproof screw or snap cover. Fix SEN1 to the bottom using double-sided tape; hold the batteries in place with rubber bands or foam rubber. Bury the container to prevent vandalism and retrieve the data later. Remember that when you go back to dig it up, the shovel will cause the last recorded event! Don't forget to turn off the power when transporting the Seismic Detector.

Ω



Testing Semiconductors

THIS MONTH WE WILL TAKE A BREAK FROM HEAVY TROUBLESHOOTING TO GO OVER SOME OF THE BASIC TECHNIQUES FOR TESTING OF COMMON SEMICONDUCTORS. WHILE FANCY TRANSISTOR TESTERS AND CURVE TRACERS HAVE THEIR

place, when you just want to know if your horizontal output transistor is toast or not they can be more trouble than they are worth. Most (though not all) failures of diodes, transistors, SCRs, and so forth, result in unambiguous changes to the device parameters that can be found with very simple tests typically using a DMM or VOM, a power supply, and a few resistors.

In most cases, these simple tests will identify bad silicon transistors. Gain, frequency response, etc., are not addressed here. While the tests can be applied to germanium devices, these are more likely to change characteristics, it would seem, without totally failing.

Safety Considerations

None of the tests described here require probing live circuits. However, before touching, probing, or unsoldering any component, make sure the equipment is unplugged and any large capacitors have been safely discharged. Not only can coming in contact with a live circuit or charged capacitor ruin your day, but your test equipment could be damaged or destroyed as well.

VOMs and DMMs

Analog and digital meters behave quite differently when testing nonlinear devices like diodes and transistors. For example, it is important to remember that an analog VOM on the lowest resis-

tance range could put out too much current for smaller devices, possibly damaging them. Ironically, this is more likely with better meters like the Simpson 260, which can test to lower ohms ($\times 1$ scale). Use the next higher resistance range in this case or a DMM, as those never drive the device under test with significant current. However, this can result in false readings as the current may be too low to adequately bias the junctions of some power devices or devices with built-in resistors.

Testing Diode Junctions with a Multimeter

On the (analog) VOM, use the low ohms scale. A regular signal diode or rectifier should read a low resistance (typically $\frac{2}{3}$ scale or a couple hundred ohms) in the forward direction and infinite (nearly) resistance in the reverse direction. It should not read near 0 ohms (shorted) or open in both directions. A germanium diode will result in a higher scale reading (lower resistance) due to its lower voltage drop.

For the VOM, you are measuring the resistance at a particular (low current) operating point—this is not the actual resistance that you will see in a power rectifier circuit, for example.

On a (digital) DMM, there will usually be a diode-test mode. Using this, a silicon diode should read between 0.5 to 0.8 V in the forward direction and open in reverse. For a germanium diode, it

will be lower, perhaps 0.2 to 0.4 V or so in the forward direction. Using the normal resistance ranges—any of them—will usually show open for any semiconductor junction since the meter does not apply enough voltage to reach the value of the forward drop. Note, however, that a defective diode may indeed indicate a resistance lower than infinity especially on the highest ohms range. So, any reading of this sort would be an indication of a bad device but the opposite is not guaranteed.

Note: For a VOM, the polarity of the probes is often reversed from what you would expect from the color coding—the red lead is negative with respect to the black one. DMMs usually have the polarity as you would expect it. Confirm this using a known diode as a reference. Also, “calibrate” your meter with both silicon and germanium semiconductors so you will know what to expect with an unknown device.

Transistor Testing Methodology

As with diode junctions, most digital meters show infinite resistance for all six

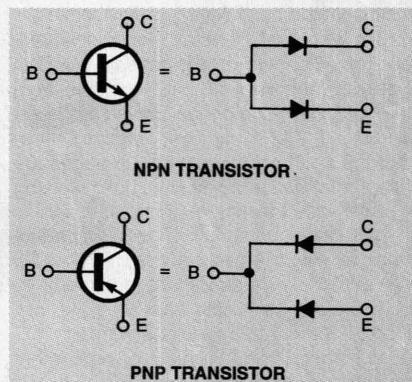


FIG. 1—A BIPOLAR TRANSISTOR can be modeled using diodes.

combinations of junction measurements since their effective resistance-test voltage is less than a diode junction drop (if you accidentally get your skin involved it will show something between 200K and 2M ohms). The best way to test transistors with a DMM is to make use of the "diode-test" function, which will be described after we talk about testing with an analog meter. Regardless of whether you test with an analog or digital meter, if you read a short circuit (0 ohms or voltage drop of 0) or the transistor fails any of the readings, it is bad and must be replaced. This discussion is for **out-of-circuit** transistors only.

One exception to this occurs with some power transistors, which have built-in diodes (damper diodes reverse connected across the C-E junction) and resistors (B-E, around 50 ohms) that will confuse these readings. If you are testing a transistor of this type—horizontal output transistors are the most common example—you will need to compare with a known good transistor or check the specifications to be sure. There are some other cases as well. So, if you get readings that do not make sense, try to confirm with a known good transistors of the same type or with a spec sheet.

Before testing an unknown device, it is best to confirm and label lead polarity (of voltage provided in resistance or diode test mode) of your meter whether it be an analog VOM or digital DMM using a known good diode (e.g., 1N4007 rectifier or 1N4148 signal diode) as discussed below. This will also show you what to expect for a reading of a forward-biased junction. If you expect any germanium devices, you should do this with a germanium diode as well (e.g., 1N34).

Note that this discussion assumes that a transistor can be tested for shorts, opens, or leakage, as though it is just a pair of connected diodes. The equivalent diode circuits for NPN and PNP bipolar transistors are shown in Fig. 1.

Obviously, simple diodes can be tested as well using this technique. However, LEDs (their forward drop is too high for most meters) and Zeners (their reverse breakdown—Zener voltage—is too large for most meters) cannot be fully tested in this manner. Those and other types of devices are discussed later in this article.

Testing With an Analog VOM

For NPN transistors, lead "A" is

black and lead "B" is red; for PNP transistors, lead "A" is red and lead "B" is black (Note: This is the standard polarity for resistance, but many multimeters have the colors reversed since this makes the internal circuitry easier to design; if the readings don't give this way, switch the leads and try it again). Start with lead "A" of your multimeter on the base and lead "B" on the emitter. You should get a reasonable low resistance reading. Depending on scale, this could be anywhere from 100 ohms to several thousand ohms. The actual value is not critical as long as it is similar to the reading you got with your "known good diode test" above. All silicon devices will produce somewhat similar readings and all germanium devices will result in similar but lower resistance readings.

Now move lead "B" to the collector. You should get nearly the same reading. Now try the other four combinations, and you should get a reading of infinite ohms (open circuit). If any of those resistances is wrong, replace the transistor. Only two of the six possible combinations should show a low resistance; none of the resistances should be near 0 ohms (shorted).

As noted above, some types of devices include built-in diodes or resistors, which can confuse these measurements.

Testing With a Digital DMM

Set your meter to the diode test. Connect the red meter lead to the base of the transistor. Connect the black meter lead to the emitter. A good NPN transistor will read a junction drop voltage of between 0.45v and 0.9v. A good PNP transistor will read open. Leave the red meter lead on the base and move the black lead to the collector. The reading should be the same as the previous test. Reverse the meter leads in your hands and repeat the test. This time, connect the black meter lead to the base of the transistor. Connect the red meter lead to the emitter. A good PNP transistor will read a junction drop voltage of between 0.45 V and 0.9 V. A good NPN transistor will read open. Leave the black meter lead on the base and move the red lead to the collector. The reading should be the same as the previous test. Place one meter lead on the collector, the other on the emitter. The meter should read open. Reverse your meter leads. The meter should read open. This is the same for both NPN and PNP transistors.

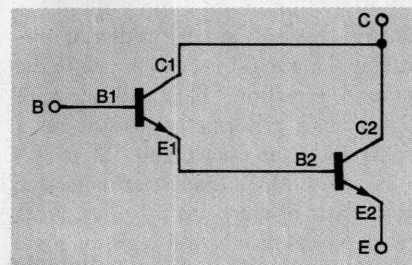


FIG. 2—A DARLINGTON TRANSISTOR is two transistors within a single package and is connected as shown here.

Again, as noted, some transistors will have built-in diodes or resistors which can confuse these readings.

Testing Power Transistors

Power transistors without internal damper diodes test just about like small signal transistors using the dual diode model, high in one direction, either B-E or B-C. If there is a built-in damper diode, it is across C-E junction and is back biased under normal operating conditions. Therefore, a reading between C-E will also test low in one direction and B-C will show a double diode drop in the reverse direction.

Also, there is often a low value resistor—about 50 ohms—between B-E when there is a built-in damper. This will show up as a nearly zero-volt junction drop on the diode test scale of a DMM, but such a reading does not indicate a bad part. Use the resistance scale to confirm.

Testing Darlington Transistors

A Darlington transistor—see Fig. 2—is a special type of configuration usually consisting of 2 transistors fabricated on the same chip or at least mounted in the same package. Discrete implementations as well as Darlington transistors with more than 2 transistors are also possible.

In many ways, a Darlington configuration behaves like a single transistor where:

- the current gains of the individual transistors it is composed of are multiplied together and,
- the B-E voltage drops of the individual transistors it is composed of are added together.

Darlington transistors are used where drive is limited and the high gain—typically over 1000—is needed. Frequency response is not usually that great, however.

Testing a Darlington with a VOM or

DMM is basically similar to that of normal bipolar transistors, except that in the forward direction, B-E will measure higher than a normal transistor on a VOM (but not open), and 1.2 to 1.4 volts on a DMM's diode-test range due to the pair of junctions in series. Note that 1.2 volts may be too high for some DMMs, and thus a good Darlington may test open—confirm that the open circuit reading on your DMM is higher than 1.4 volts or check with a known good Darlington.

Testing Digital or Bias Resistor Transistors

Occasionally you may find a transistor that includes an internal bias resistor network attached to the base and emitter so that it can be driven directly from a digital (e.g., TTL) source; see Fig. 3. These may be used in consumer elec-

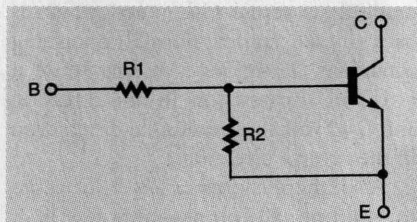


FIG. 3—AN INTERNAL BIAS RESISTOR is sometimes present to allow a transistor to be driven directly from a TTL-level signal.

tronics equipment where space is critical, or for no good reason other than to make it difficult to locate a suitable replacement device!

The addition of R1 makes testing with a multimeter other than for shorts more difficult. With a VOM, you should see a difference in the B-E and B-C junctions in the forward and reverse directions. However, a DMM will probably read open across all pairs of terminals.

Testing Zener Diodes

The following applies to both testing of Zeners for failure and determining the ratings of an unknown device:

With a VOM, a good Zener diode should read like a normal diode in the forward direction and open in the reverse direction unless the VOM applies more than the Zener voltage for the device. A DMM on its diode test range may read the actual Zener voltage if it is very low (e.g., a couple of volts) but will read open otherwise. The most common failure would be for the device to short—read 0.0 ohms in both direc-

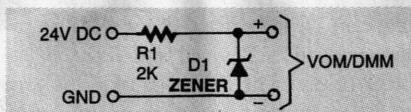


FIG. 4—USE THIS CIRCUIT to test Zener diodes.

tions. Then, it is definitely dead!

Some Zeners are marked with a JEDEC (1N) part number, others with a couple of colored bands (e.g., for 18 volts the bands would be brown and gray), or a house number or house color code.

You can easily test a Zener and identify its voltage rating with a DC power supply, resistor, and multimeter. You will need a power supply (a DC wall adapter or AC wall adapter with a rectifier and filter capacitor is fine) greater than the highest Zener voltage you want to test. Select a resistor that will limit current to a few mA. For example, for Zeners up to about 20 V, you can use the circuit in Fig. 4. This same approach applies to other devices that exhibit a similar behavior, such as the B-E junction of a bipolar transistor.

Testing Unijunction and Programmable Unijunction Transistors

Unijunction Transistors (UJTs) and Programmable Unijunction Transistors (PUTs) are used in similar sorts of circuits, though the UJT is all but extinct. They both exhibit a negative resistance characteristic and can be used easily in low to medium frequency free-running relaxation oscillators and other trigger type circuits.

The UJT (see Fig. 5) goes into heavy conduction from E to B1 when E becomes more positive than a critical trigger voltage, $V_T = n \times V_{BB} + 0.6$. (Note, n , the "intrinsic standoff voltage," is typically about 0.6.) It continues to conduct until the emitter current drops

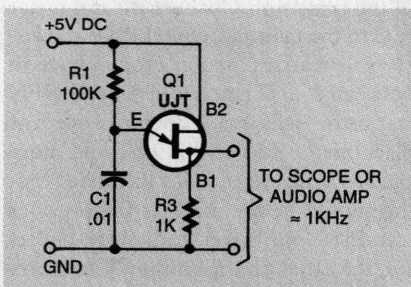


FIG. 5—UNIUNCTION TRANSISTORS are not that common these days, but if you come across one this circuit can be used to test it.

below some minimum 'valley current' value. (Sounds sort of like a thyristor, right?)

The PUT (see Fig. 6) is even more like a thyristor in that the triggering takes place when the G terminal becomes more positive than the A (probably plus a diode drop, 0.6 V) so that the threshold voltage can now be set with a voltage divider feeding the anode.

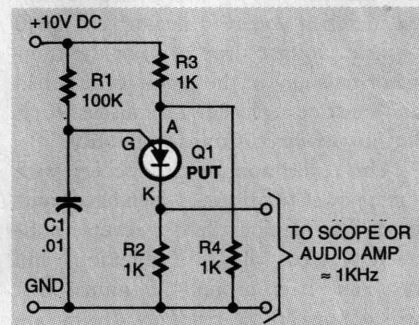


FIG. 6—A CIRCUIT FOR TESTING programmable unijunction transistors (PUTs) is shown here.

Then, current flows from the G to the K terminal. Note that its leads are even labeled like an SCR, but it behaves sort of backwards!

For an initial test, check between B1 and B2 (UJT) or A and K (PUT) with an ohmmeter. The resistance should be the same in both directions and typically a few thousand ohms or more. A short or wildly different readings would indicate a bad device. This doesn't prove that the device is good—only that it hasn't blown up. A more complete test requires a simple circuit and some means of detecting an audio output signal. For the UJT, use the set up in Fig. 5; for the PUT, an additional voltage divider (R3 and R4) is needed to set the threshold as shown in Fig. 6.

Testing SCRs and Triacs

Next, let's deal with testing various types of thyristors. For SCRs, the gate to cathode should test like a diode (which it is) on a multimeter. The anode to cathode and gate to anode junctions should read open. For Triacs, the gate to main terminal 2 (MT2) should test like a diode junction in both directions. MT1 to MT2 and the gate to MT1 junctions should read open. For diacs and sidacs, there is no gate terminal—resistance should be infinite in both directions.

Note: Some thyristors will have a low G-K/MT2 resistance but it should not read as a short.

The real test is quite simple but will

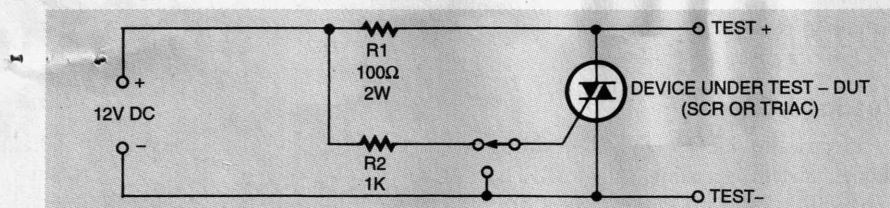


FIG. 7—IN THIS SCR/TRIAC test circuit, R1 limits current to the device under test, and R2 limits current to its gate.

require a low voltage DC power supply and two resistors. For Triacs, a negative output from the supply is also desirable to test the triggering when the gate is negative.

To test SCRs or Triacs, use the circuit shown in Fig. 7. There, R1 is used to limit current through the device and R2 is used to limit current to the gate. The 12-VDC supply should be capable of currents to at least 200 mA. Note that R1 calls for a unit with a 2-watt rating; a $\frac{1}{2}$ -watt unit should be sufficient for R2. Those ratings should work for testing most small to medium power devices. Check the "minimum gate current" and "holding current" specs to be sure. For larger devices, R1 and/or R2 might need to be smaller.

Here's the procedure:

1. Connect the supply as shown.
2. Trigger the gate from the positive of the supply through the current limiting resistor (R2) and see that the DUT turns on and stays on when the gate is disconnected.
3. Open the circuit to the anode (with the gate connected to the cathode) and again reconnect the anode resistor. The DUT should now be off again.
4. For Triacs, repeat steps (2) and (3) with R2 supplied from a negative voltage.

If the device passes these tests, it is behaving properly and is probably functional. However, without applying full voltage or current, there is no way of knowing if it will meet all specifications.

You can replace the DC supply with a low voltage power transformer (say, 12 VAC). Use a scope to monitor the voltage across the DUT or R1. Then, when the gate is connected to R2, you should see the voltage across the DUT drop to nearly zero when it switches on partway through the positive cycle. This phase will be determined by the voltage and value of R2. It should remain off for the entire negative cycle (SCRs only) with the gate connected and remain off all the time with the gate connected to the

cathode.

Testing Diacs and Sidacs

Diacs and Sidacs are thyristors without a gate terminal. They depend on the leakage current to switch them on once the voltage across the device exceeds their specified ratings. With an ohmmeter, they can be tested only for shorts. Resistance should be infinite in both directions.

However, you can test a diac or sidac with a resistor, variable power supply (you will need at least the rating of the device), and a DMM. Hook them in series as shown in Fig. 8 and monitor across the device. With care, your variable supply can be a Variac, a 1N4007 diode, and 1 μ F, 200 V capacitor as shown. The 47,000-ohm resistor (R1) is used to limit the current:

CAUTION: This circuit is not isolated from the power line. I recommend

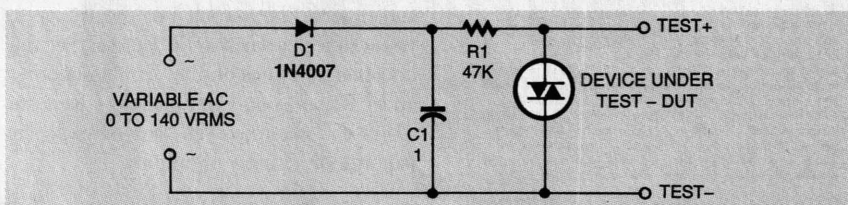


FIG. 8—A DIAC OR SIDAC is a thyristor without a gate. Use this circuit to test one.

using an isolation transformer between the variable voltage source (Variac) and the power line for safety. If the DUT is rated at more than about 180 volts, you will need to use a doubler and higher-voltage capacitor, but testing is otherwise similar.

As you increase the input, the voltage on the DUT will track it until the rated voltage, at which point it will drop abruptly to zero and stay there until the voltage is reduced below its holding current. Repeat with the opposite polarity.

With a scope, testing is even easier as you can use an AC supply directly (remove D1 and C1) and observe that the DUT will turn on at the proper voltage on both polarities of the AC wave-

form and stay on until the voltage crosses 0. But, again, use an isolation transformer for safety.

Identifying Unknown Bipolar Transistors

The type (NPN or PNP) and lead arrangement of unmarked transistors can be determined using a multimeter and the back-to-back diode model discussed earlier. The collector and emitter can be identified based on the fact that the doping for the B-E junction is always much higher than for the B-C junction. Therefore, the forward voltage drop will be very slightly higher—this will show up as a couple of millivolts (sometimes more) difference on a DMM's diode-test scale or a slightly higher resistance on an analog VOM.

To determine the lead arrangement, label the pins on the unknown device 1, 2, and 3. Put the positive probe (as determined previously) of your multimeter on pin 1. Now, measure the resistance (VOM) or diode drop (DMM) to the other two pins. If the positive probe is on the base of a good NPN transistor, you should get low resistance readings or a low diode drop to the other two leads. The B-C resistance or diode drop will be just slightly lower than the B-E reading.

If one or both measurements to the

other two pins is high, put the positive probe on pin 2 and try again. If still no cigar, try pin 3.

If this still doesn't work, you may have a PNP transistor—repeat with the negative probe as the common pin.

If none of the six combinations yields a pair of low readings—or if more than one combination results in a pair of low readings, your transistor is likely bad—or it is not a bipolar transistor at all!

As noted, some power transistors have built-in base resistors or damper diodes and will confuse these measurements. However, the lead arrangement of these types of transistors is usually self-evident (standard TO3, TOP3, or TO220 cases). There are also some tran-

sistors with series base resistors that may prove confusing. These are relatively rare, however.

Testing MOSFETs

Verify that the gate has infinite resistance to both drain and source. There is one exception to that: FETs with protection circuitry may act like there is a Zener shunting the gate and source terminals (*i.e.*, diode drop for gate reverse bias, about 20-volts breakdown in forward bias).

Connect the gate to the source. The drain to source junction should act like a diode.

Forward bias the gate-source junction with about 5 volts. The drain-source junction in forward bias should measure very low resistance. In reverse bias, it will still act like a diode.

The usual failure mode you are likely to see with these devices is both a gate-source and drain-source short. In other words, everything connected together. Also, when

handling these devices remember that MOSFETs are static sensitive!

Testing LEDs

Electrically, LEDs (and IR-emitting diodes, strictly speaking called IREDs) behave like ordinary diodes except that their forward voltage drop is higher. Typical values are IR: 1.2 volts; Red: 1.85 volts; Yellow: 2 volts; and Green: 2.15 volts. The new blue LEDs will be somewhat higher (perhaps 3 volts). These voltages are at reasonable forward current. Depending on the actual technology (*i.e.*, compounds like GaAsP, GaP, GaAsP/GaP, GaAlAs, etc.), actual voltages can vary quite a bit. For example, the forward voltage drop of red LEDs may range at least from 1.5 to 2.1 volts. Therefore, LED voltage drop is not a reliable test of color, though multiple samples of similar LEDs should be very close. Obviously, if the device is good, it will also be emitting light when driven in this way if the current is high enough.

You can test for shorts and opens with a multimeter (but it must be able to supply more than the forward voltage drop to show a non-open condition). However, an LED can be weak and still pass the electrical tests, so checking for output is still necessary. Therefore, even if these tests don't find a problem, drive the LED from a DC supply and appropriate current-limiting resistor and observe the output. For the IR types, you will need a suitable IR detector. See the document: "Notes on the Troubleshooting and Repair of Hand-Held Remote Controls" on my Web site (www.repairfaq.org) for a variety of options.

Testing Optoisolators and Photo-Interrupters

Both of these classes of components are basically similar: a light source (usually an IR LED) and photodetector together in a single package. The optoisolator will be totally sealed with adequate separation between the two parts to provide the specified isolation voltage rating. The photo-interrupter (and similar devices) will provide a beam path that can be blocked or otherwise modified by external means.

For both types, the photodetector can be a photodiode, phototransistor, photothyristor, or another more complex device or circuit. Refer to an optoelectronics databook or the catalog of a large electronics distributor for specific pinouts and specifications.

Assuming a photodiode or photo-

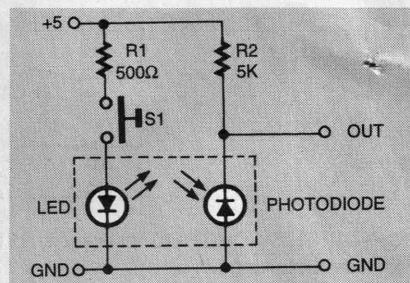


FIG. 9—A SIMPLE CIRCUIT for testing opto-isolators.

transistor type (most common), these can be tested for basic functionality pretty easily. Wire up the test circuit shown in Fig. 9. Depressing S1 should result in the output dropping from +5 volts to close to 0. For monitoring on a scope, drive the LED with a pulse generator and current-limiting resistor instead of S1. With a photo-interrupter type, blocking or adding a reflector to the optical path (as appropriate) should result in similar behavior.

Testing Thermistors

There are two types of thermistors: Positive Temperature Coefficient (PTC) types have a resistance that increases with increasing temperature. Negative Temperature Coefficient (NTC) types have a resistance that decreases with increasing temperature.

For a small thermistor, put an ohmmeter on it and heat the device under test with a blow dryer, heat gun, or the tip of a soldering iron—the resistance should change smoothly (up or down depending on whether it is PTC or NTC type). If the resistance changes erratically or goes to infinity or zero, the device is bad. However, you will need specifications, temperature measuring sensors, etc., to really determine if it is operating correctly.

Wrap-Up

Next time, we will continue our discussion of semiconductor testing with additional service-related information and some very simple curve tracer and in-circuit tester schematics. In a future column, we'll discuss the specific problem of testing the laser diodes found in CD and DVD equipment and laser printers. Until then, check out my Web site: www.repairfaq.org. I welcome comments (via e-mail please at sam@stdaids.picker.com) of all types and will reply promptly to requests for information. See you next time!

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Q & A

READERS' QUESTIONS, EDITORS' ANSWERS
CONDUCTED BY MICHAEL A. COVINGTON, N4TMI

One-RPM Stepper

Q For an astronomical instrument, I need to make a stepper motor turn at exactly 1 revolution per minute (rpm). The motor is a Mitsumi M68SP-4 unit with six wires. It moves in 1.8-degree steps. How do I do this?—M. McA., Tucson, AZ

A Your letter explains that you're building a "barn-door tracker," a simple gadget that lets you take long-exposure photographs of the stars by compensating for the earth's rotation. It consists of two pieces of wood, hinged together, with the axle of the hinges pointed toward Polaris and therefore parallel to the earth's axis. The hinges gradually swing open at an angular rate equal to the earth's rotation as the wooden panels are pushed apart by 20-thread-per-inch screw rotating at 1 rpm and positioned exactly 290mm from the hinge axle. Naturally, the device can only open up a few degrees before you have to reset the screw, but that's enough for a time exposure of five or ten minutes. For details, see *Astrophotography for the Amateur*, by Michael Covington, 1999 edition, pp. 120-121, or do a Web search for "barn-door tracker" or "Scotch mount."

Most people turn the handle of a barn-door tracker by hand, but you've elected to use a stepper motor. Accordingly, this is a two-part question: how to run a stepper motor, and how to get the desired speed. Let's tackle the second part first.

A stepper motor turns when its windings are energized in a specific sequence, and by far the easiest way to do this is to use a stepper motor driver chip such as the Motorola MC3479P (see Fig. 1). This chip moves the motor in either full or half steps depending on whether pin 9 is connected to V+ or to ground. Each step is triggered in turn by a clock pulse on pin 7.

Since your motor makes 1.8-degree

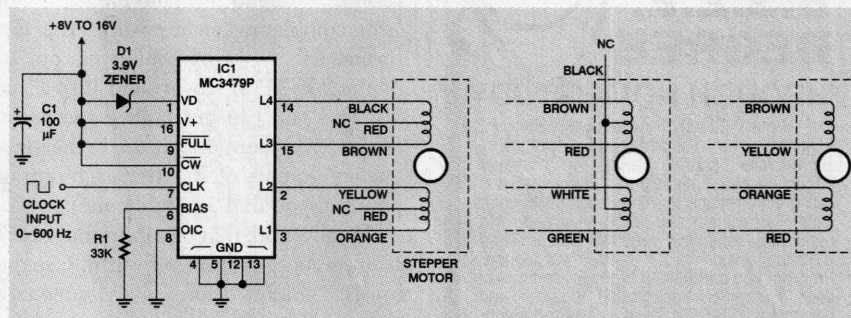


FIG. 1—MOTOROLA'S MC3479P CAN BE USED to drive most 4-, 5-, and 6-wire stepper motors with a minimum of additional components. To be on the safe side, don't assume your motor's wire colors match what's shown here; check things out with an ohmmeter to be sure.

steps, it needs 200 pulses (or, in half-step mode, 400 pulses) to make a complete circle. A frequency of 400 cycles per minute equals 6.667 cycles per second (Hertz). Accordingly, the second part of the task is to generate a precise 6.667-Hz clock signal.

That was the challenging part. The easy way would be to program a microcontroller to count the appropriate number of microseconds for each cycle. However, you requested a design using off-the-shelf parts, and after trying lots of combinations I came up with the quartz-controlled circuit in Fig. 2. Here a standard 4.9152-MHz microprocessor crystal is used with a CD4060 (MC14060) CMOS oscillator-divider, which divides it down to 600 Hz. Then

two CD4017 (MC14017) decade counters divide the frequency by 9 and by 10, respectively. The output is a 6.667-Hz square wave.

Several cautionary notes: In the oscillator-divider circuit, don't use 74HC chips (74HC4060, 74HC4017) unless you're using a regulated 5-volt supply. In the motor circuit, the MC3479P can handle a maximum of 350 mA; for heavier currents than that, you need a similar but slightly different chip, the SAA1042A, described in this column in August 1996.

The color code for stepper-motor wires seems to vary haphazardly, so use an ohmmeter to check out your particular motor before hooking it up. If the windings have center taps (yours do),

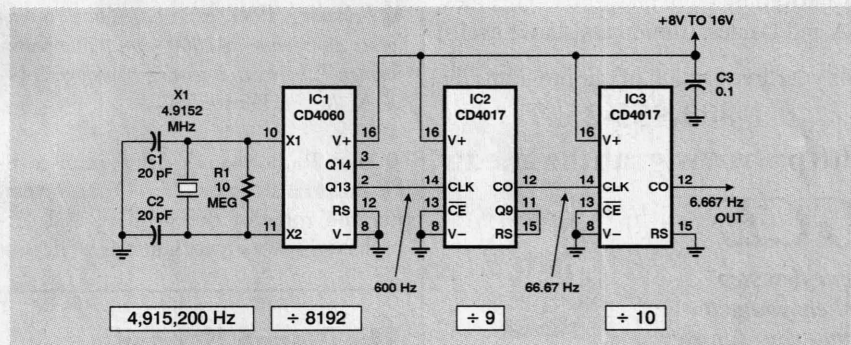


FIG. 2—THIS CIRCUIT GENERATES a precise 6.667-Hz clock signal using off-the-shelf components. The 4.1952-MHz crystal (X1) is used in many microprocessor circuits and is easy to find.

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leave them unconnected. If the motor turns the wrong way, reverse the connections to one of the windings. With a motor that moves in 3.6-degree steps (100 steps or 200 half-steps per full circle), you'll need to cut the clock speed in half; you can do this by feeding IC2 from pin 3 of IC1 instead of pin 2.

You may want to implement a "fast reverse" mode for backing the screw out after completing an exposure. You can reverse the motor by connecting pin 10 of the MC3479P to ground instead of V+, and you can get higher speeds by feeding IC2 from one of the higher-frequency outputs of IC1 (try each output pin until you find a suitable one).

Finally, note that in continuous operation, stepper motors get warm. Use the lowest voltage that gives adequate torque. You can reduce the voltage to the motor by including some silicon rectifiers (1N4001 or equivalent) in series with the power supply; each one takes off about 0.7 volt, and they also protect against an accidentally reversed battery.

The MC3479P and some motors that work well with it are available from Jameco, 1355 Shoreway Road, Belmont, CA 94002; Tel: 800-831-4242; Web: www.jameco.com. Suitable motors are

very common in junked printers and the like; it may be that all you need is the chip. The ECG1857 and NTE1857 are exact substitutes for the MC3479P.

To learn more about stepper motors, see "Robotics Workshop" in **Popular Electronics**, July 1999, pp. 64—66, 70 (available from our Reprint Bookstore) and Doug Jones' stepper motor FAQ on the Web at www.cs.uiowa.edu/~jones/step.

Audio Transformer Query

Q I'd like to build the intercom in your February, 1997, issue, p. 14, but I cannot locate an 8-ohm-to-1000-ohm audio transformer. Where can I get this transformer?—G. N. K., Ft. Worth, TX

A Use RadioShack's 1000-ohm-to-8-ohm transformer (273-1380), reversing the roles of the primary and secondary. That will work just as well.

AM Stereo Wanted

Q Is there a circuit or a chip that will add AM stereo capability to an existing AM radio?—R. A. H., Lancaster, OH

A Somehow, AM stereo has never been the commercial success that was predicted, perhaps because the talk-radio boom came along at about the same time. But it hasn't died out, either.

We published an AM stereo conversion circuit in our January 1984 issue, pages 41—46, 102, and 114 (see also the May 1984 issue, p. 20). It used a Motorola MC13020P chip, which is going out of production and may be hard to find.

You can buy a wide range of AM stereo receivers and tuners from AM Stereo Works (1555 State Road 207, St. Augustine, FL 32086; Tel: 904-810-5140; Web: www.stereoam.com). They have conversion kits for existing receivers, but they are fairly expensive (about \$90); they can also convert your radio for you.

Pager Transmitter?

Q Is there a way to rig a 9-volt transmitter to directly activate my pager (without going through the telephone lines or my pager company) when a motion detector is triggered?—H.E.R., Shell Beach, CA

A It doesn't sound easy. You'd have to mimic the activation code sent to your pager by the pager company, which may be a fairly complicated signal. Consider using a separate receiver (such as a pocket FM radio) and building one of the FM transmitters in **Popular Electronics**, July 1999, pp. 37—42. On using your PC to generate the digital codes used by pagers, see **Electronics Now**, February 1998 issue, pp. 42—45.

Parallel 4049s; Another Chip Quest

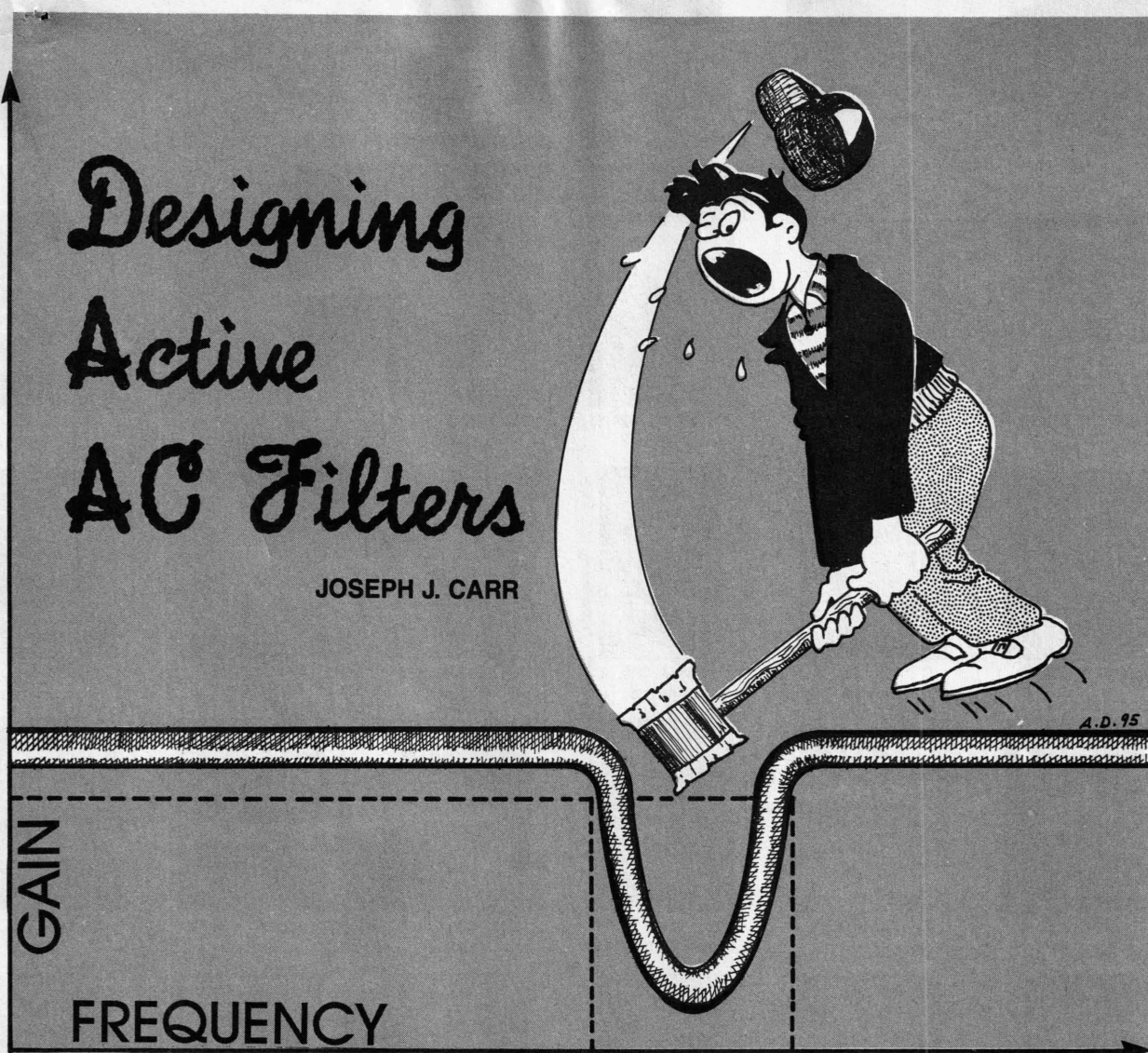
Q Is it possible to connect two or more elements of the MC14049 (CD4049) buffer/inverter chip to increase the drive and sink capability?

Also, I would appreciate any information you can give me about the BTS629 7-pin automotive dimmer chip.—S. B., Mississauga, Ont., Canada

A Your first question is easy: Yes, CMOS inverters (including units of the 4049) can be paralleled to increase their output current capacity. Connect

Designing Active AC Filters

JOSEPH J. CARR



Low-pass, high-pass, band-pass and notch—they're all here!

ELECTRONIC FREQUENCY-SELECTIVE AC filters (*filters*) are circuits that favor some frequencies and discriminate against other frequencies. Frequencies that pass through the filter with little attenuation are the *passband*, while attenuated frequencies are the *stopband*.

Filter circuits are classified in several ways: Passive vs. active, analog vs. digital vs. software, by frequency range (e.g. audio, RF, or microwave), or by passband characteristic. *Passive filters* are made of various combinations of passive components such as resistors (R), capacitors (C), and inductors (L). In general, passive filters are lossy; they attenuate the signal

and amplification might be ultimately required. *Active filters*, on the other hand, are based on active electronic devices such as a transistor or an operational amplifier along with passive components (R, C, and L) that determine frequency. The passive components are customarily composed of resistors and capacitors for they are cheaper and easier to obtain than inductors.

There are all kinds of filters. *Active filters* use analog linear-circuit techniques such as those applying to operational amplifiers. *Digital filters* use digital IC devices, and are often based on capacitor-switching techniques. *Software filters* im-

plement solutions to frequency-selective equations using computer programming techniques.

Filters are also classified by frequency range. *Audio filters* operate from the sub-audio to ultrasonic ranges (near-DC to about 20-kHz). *RF filters* operate at frequencies above 20-kHz, up to about 900-MHz. *Microwave filters* operate at frequencies greater than 900-MHz. These range designations are not absolute, but do serve to indicate approximate points at which a change of design technique generally takes place. The filter circuits discussed in this article are audio-active filters that have a passband between

the sub-audio and the low-ultrasonic regions.

Finally, filters are classified by the nature of their frequency-response characteristics. This method of categorizing filters takes note of the filter's passband and stopband. Low-pass filters, high-pass filters, band-pass filters, and stopband filters will be discussed in this article.

Filter characteristics

Figure 1 shows the frequency-range categories for *ideal* filters. A low-pass filter has a passband from DC (zero Hertz) to a specified cut-off frequency (f_1). All frequencies above the cut-off frequency are attenuated, so they are in the stopband. A band-pass filter has a passband between a lower limit (f_2) and an upper limit (f_3). All frequencies lower than f_2 or greater than f_3 are in the stopband. A high-pass filter has a stopband from DC to a certain lower limit (f_4). All frequencies greater than f_4 are in the passband.

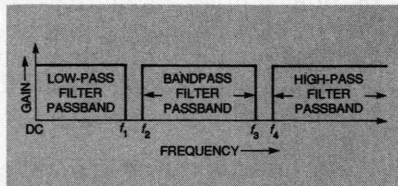


FIG. 1—IDEALIZED frequency-responses curves for low-pass, bandpass and high-pass AC filter circuits.

A stopband filter response is shown in Fig. 2. This filter attenuates frequencies between lower and upper limits (f_5 to f_6), but passes all others. When the stopband is very narrow, the stopband filter is called a notch filter. This filter type is often used to remove a single, unwanted frequency. An example of such an application is removal of unwanted 60-Hz interference (AC hum) caused by proximity to the AC power line.

Ideal vs. practice

The frequency-response

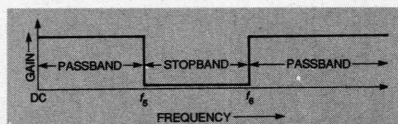


FIG. 2—IDEALIZED frequency-response curve for stopband AC filter circuits.

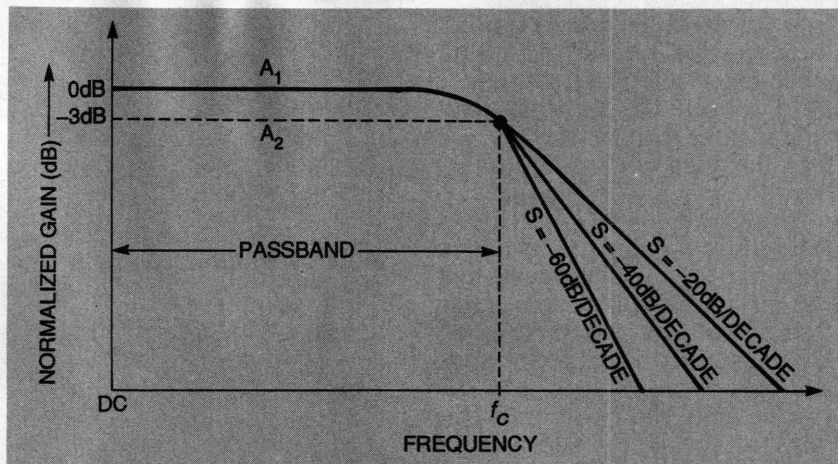


FIG. 3—TYPICAL LOW-PASS FILTER passband frequency-response curves for first, second and third-order Butterworth filters.

curves shown in Figs. 1 and 2 are idealized. In real life the corners of the curves are rounded and not squares, and the vertical lines are gentle to steep slopes as illustrated in Fig. 3. The Butterworth, Chebyshev, Cauer (also sometimes called elliptic), and Bessel filters are among the typical filters used. The ideal Butterworth frequency-response curve is shown in Fig. 3. The noteworthy properties of the Butterworth filter is that both the passband and stopband are relatively flat, and the transition region slope between them is continuous.

Standard practice in filter nomenclature is to specify the passband between frequencies where the frequency response falls off 3 dB from the mid-passband gain. In Fig. 3, the cut-off frequency (f_c) is at the point where gain (A_2) falls off by a factor of 0.707 to a designated gain value called A_1 . At that spot in the frequency-response curve, the reference point is called the -3-dB point .

At frequencies greater than f_c , the gain falls off linearly at a rate that depends on the order of the filter. The slope (S) of the fall off is measured in either decibels per octave (a 2:1 frequency change) or decibels per decade (a 10:1 frequency change). Note that these two specifications can be scaled relative to each other: -6 dB/octave has the same slope as -20 dB/decade . The slopes shown in Fig. 3 cover three Butterworth cases. A first-

order filter has a roll-off of -20 dB/decade , a second-order filter has a roll-off of -40 dB/decade , and a third-order filter has a roll-off at -60 dB/decade . These slopes are the same as 6, 12 and 18-dB/octave, respectively.

It might appear that only third-order filters are used because the transition from passband to stopband is most rapid (steep). But higher-order frequency-response filtering is obtained at the cost of more complexity, greater sensitivity to component value error, and difficult design configurations. Some higher-order filter designs are also more likely to oscillate than lower-order equivalents. The selection of filter order is a trade-off between system needs and complexity.

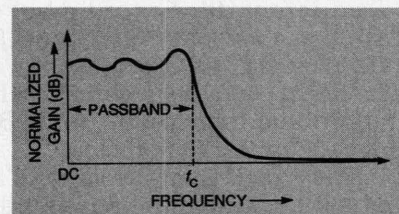


FIG. 4—PASSBAND FOR LOW-PASS Chebyshev AC filter reveals a rapid gain fall-off at the cut-off frequency, f_c .

The steepness and shape of the roll-off curve is a function of the filter's damping factor. Butterworth filters tend to be heavily damped, which explains the gradual roll-off in the response curve. The Chebyshev filter frequency-response curve shown in Fig. 4 is lightly damped, so it

has a variation (or "ripple") within the passband. The Chebyshev filter offers a faster gain roll-off than the Butterworth filter, but at the cost of less flatness within the passband.

The Caer or elliptic filter's frequency-response curve shown in Fig. 5 offers the fastest roll-off at the cut-off frequency (f_c), as well as relatively good flatness within the passband. Notches of -40 dB to -60 dB can be achieved close to f_c , but only at a cost of less attenuation further into the stopband. A typical Caer filter response has a deep notch close to f_c , rises to a peak at some high frequency, and then gradually falls off for even higher frequencies at a rate of -20 -dB to -40 -dB/decade.

The frequency-response curve for the Bessel filter is shown in Fig. 6. Although it appears similar to the Butterworth response, its gain is not flat within the passband. The benefit of the Bessel filter is a flat phase response across the passband.

Filter phase response

Most filter circuits exhibit a phase change over the passband. The responses for the Butterworth and Bessel filters are shown in Fig. 7. Note that the maximal flat Butterworth curve exhibits a decidedly non-linear phase response in both passband and stopband. The frequency dependent phase shift of a low-pass filter is -45 degrees at f_c , and increases by a factor of -45 degrees for each additional increase of -20 -dB/decade in the roll-off slope.

The Bessel filter also shows a phase shift over the passband, but it is nearly linear. A useful feature of this characteristic is that it allows a uniform time delay all across the passband. As a result, the Bessel filter offers the ability to pass transient pulse wave forms with minimum distortion. The Bessel filter is said to work best at the frequency where $f = f_c/2$.

Low-pass filters

A diagram for a low-pass filter

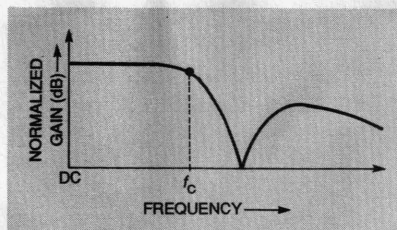


FIG. 5—PASSBAND for the Caer or Elliptic AC filter.

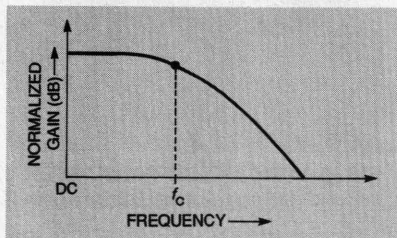


FIG. 6—PASSBAND for the Bessel AC filter.

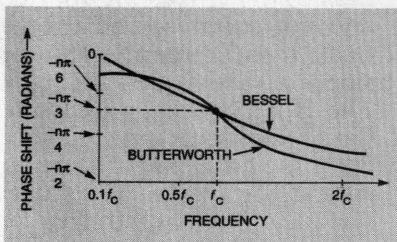


FIG. 7—COMPARISON of signal phase response for the Butterworth and Bessel AC filters.

circuit is shown in Fig. 8. This filter is called the voltage-controlled-voltage-source (VCVS) filter. The basic configuration is a non-inverting-follower operational amplifier (IC1). The operational amplifier selected should have a high-gain bandwidth, relative to the cut-off frequency, in order to permit the filter to operate properly. The gain of the circuit is given by:

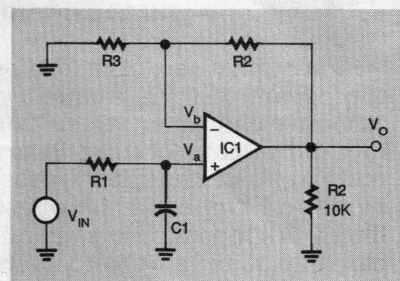


FIG. 9—FIRST-ORDER VCVS low-pass filter circuit using a typical operational amplifier.

$$A_v = (R_f/R_{in})$$

When $R_{in} = R$:

$$R_f = R(A_v - 1)$$

In some circuits, the gain may be unity. In those circuits the resistor voltage-divider feedback network is replaced with a single connection between output and the inverting input.

The input circuitry of the generic VCVS filter consists of a network of impedances labeled Z1 through Z6. Each of these blocks will be either a resistance (R) or a complex capacitive reactance ($-jX_c$). Which element becomes which type of component is determined by whether the filter is a low-pass or high-pass type.

The order of the filter, denoted by n , refers to the number of poles in the design, or in practical terms, the number of RC sections. A first-order filter ($n = 1$) consists of Z3 and Z6 (Fig. 8), a second-order filter ($n = 2$) consists of Z2, Z3, Z5 and Z6, and a third-order filter ($n = 3$) consists of all six impedances (Z1 through Z6). Higher order filters (n greater than 3) can also

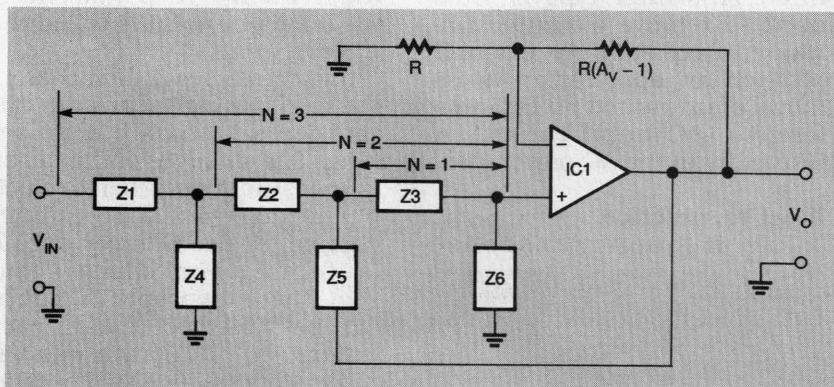


FIG. 8—HERE'S A SIMPLIFIED SCHEMATIC for a low-pass design known as a voltage-controlled-voltage-source (VCVS) filter.

be built. In a low-pass filter Z1 through Z3 are resistances, while Z4 through Z6 are capacitances. The component types are reversed in high-pass filters.

First-order low-pass filters

The first-order low-pass VCVS filter is shown in Fig. 9, and its frequency-response curve is shown in Fig. 3 (slope $S = -20$ -dB/decade). The filter consists of a single-section RC low-pass filter driving the non-inverting input of an operational amplifier. The gain of the operational amplifier is $[(R2/R3) + 1]$. The high-input impedance of IC1 prevents loading of the RC network. The general form of the transfer equation for the gain vs. frequency response for the first-order filter is:

$$A_{dB} = 20\text{LOG}(A_v) - 20\text{LOG}[1 + (\omega/\omega_c)^2]^{1/2}$$

where: A_{dB} is the gain of the circuit in decibels; A_v is the voltage gain within the passband; LOG denotes the base-10 logarithms; ω_c is the ratio of the input frequency to the cut-off frequency ($\omega_c = f/f_c$).

The voltage at the output of the RC network (V_a) is found from the voltage divider equation:

$$V_a = -jX_c V_{in} / (R - jX_c)$$

where: $-jX_c = 1/(j2\pi fC)$ and j is the imaginary operator $[(-1)^{1/2}]$.

Substituting the value for $-jX_c$:

$$V_a = [V_{in}/j2\pi fC]/R + [1/j2\pi fC]$$

which simplifies to:

$$V_a = V_{in} / [1 + \pi fCR]$$

If the transfer function of the non-inverting follower is:

$$V_o = V_{in} [(R2/R3) + 1]$$

and since $V_{in} = V_a$:

$$V_o = [V_{in} / (1 + 2\pi fCR)] \times [(R2/R3) + 1]$$

The above equation is put into a more workable form:

$$V_o/V_{in} = [A_v / (1 + jf/f_c)]$$

where: V_o is the output signal voltage; V_{in} is the input signal voltage; A_v is the passband gain $[(R2/R3) + 1]$; f is the signal frequency; and f_c is the -3 -dB fre-

quency ($1/2\pi RC$).

The filter parameters are required to define the operation of any particular circuit. The *gain magnitude* and *phase shift* are found from the following equations:

Gain magnitude:

$$V_o/V_{in} = A_v / [1 + (f/f_c)^2]^{1/2}$$

and *phase-shift angle* (in radians):

$$\phi = -\text{Tan}^{-1}(f/f_c)$$

Because the filter characterization depends in part on the ratio f/f_c , the equations take different forms at different values of f and f_c . These can be reduced as follows:

At low frequencies that are well within the passband (f is less than f_c):

$$V_o/V_{in} = A_v = (R2/R3) + 1$$

At the -3 -dB cut-off frequency ($f = f_c$):

$$V_o/V_{in} = 0.707A_v$$

At a high frequency that is well above the -3 -dB cut-off frequency (f greater than f_c):

$$V_o/V_{in} < A_v$$

Table 1 shows the characteristics of first-order filters at several different ratios of f/f_c .

Designing a first-order low-pass filter

There are two basic ways to

determine the component values for the low-pass filter shown in Fig. 9: ground-up and frequency-scaling methods. The following is the ground-up method:

1. Select the -3 -dB cut-off frequency (f_c) required.
2. Select a standard-value capacitor value.
3. Calculate the required resistance for $R1$:

$$R1 = 1/2\pi f_c C$$

4. Select the passband gain for f less than f_c .
5. Select a value for R , and
6. Calculate R_f from:

$$R_f = R(A_v - 1)$$

Example

A low-pass filter shown in Fig. 9 is needed to process a transducer signal. The cut-off frequency should be 100-Hz, and the gain should be 5. The solution is:

1. $F_c = 100$ -Hz
2. Select trial value for $C1$: $0.1\text{-}\mu\text{F}$
3. Calculate $R1$:

$$R1 = 1/2\pi F_c C1$$

$$R1 = 15,923 \text{ ohms}$$

4. Select a trial value for $R1$: 10,000 ohms. Then calculate R_f :

$$R_f = 40,000 \text{ ohms}$$

Design by normalized model

The filter design can be simplified by using a normalized

TABLE 1—FILTER CHARACTERISTICS

Applied Frequency (Hz)	Gain Magnitude		Phase Shift (degrees)
	A_v	dB	
10	2.00	6.020	-0.57
20	1.999	6.018	-1.15
50	1.998	6.009	-2.86
80	1.994	5.993	-4.57
100	1.990	5.977	-5.71
200	1.961	5.850	-11.31
500	1.789	5.052	-26.57
800	1.561	3.872	-38.66
1000	1.414	3.010	-45.00
2000	0.894	-0.969	-63.43
5000	0.392	-8.129	-78.69
8000	0.248	-12.109	-82.87
10000	0.199	-14.023	-84.29
20000	0.100	-20.011	-87.14
50000	0.040	-27.960	-88.85
80000	0.025	-32.042	-89.28
100000	0.020	-33.980	-89.43

model. First, design a filter for a standardized frequency (e.g. 1-Hz, 10-Hz, 100-Hz, 1000-Hz or 10,000-Hz) and list the component values. The values for any other frequency can then be computed by a simple ratio and proportion. An example of a first-order low-pass Butterworth filter is shown in Fig. 10 with the component values normalized for 1 kHz. The actual required component values ($R1'$ and $C1'$) are found by dividing the normalized values shown by the desired cut-off frequency in kilohertz:

$$C1' = (C1)(1\text{-kHz})/f$$

or

$$R1' = (R1)(1\text{-kHz})/f$$

Note: The value for f in above two equations is in kilohertz (kHz) units.

Leave one of the values alone, and calculate the other. In general, it is easier to obtain precision resistors in unusual values (or the value obtained by a potentiometer), so it is common practice to select a standard capacitance and calculate the required resistance.

Example:

Change the frequency of the normalized 1-kHz filter to 60 Hz (i.e. 0.06 kHz).

$$C1' = (C1)(1\text{-kHz})/f = 0.265 \mu\text{F}$$

Second-order low-pass filters

The circuit for a second-order, low-pass filter is shown in Fig. 11, while the response curve with a -40-dB/decade slope is shown in Fig. 3. This circuit is similar to the first-order filter (Fig. 10), but with an additional RC network in the frequency-selective portion of the circuit. The circuit is wired for unity gain.

The purpose of $R3$ in Fig. 11 is to help counteract the DC offset at the output of the operational amplifier that is created by input bias currents charging the capacitors in the frequency selective network. The value of $R3$ in the unity gain case is $2R$, where R is the value of the resistors in the frequency selective network. In cases where DC

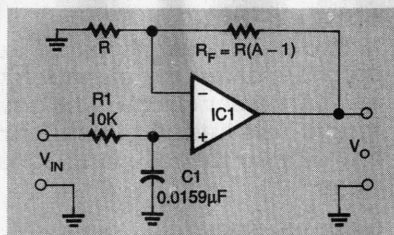


FIG. 10—FIRST-ORDER VCVS low-pass filter circuit normalized to 1-kHz by the selection of component values.

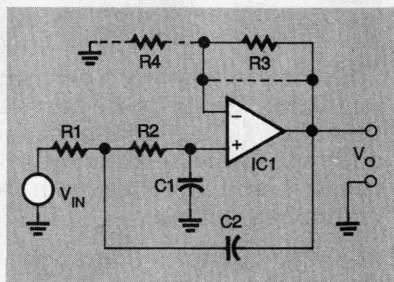


FIG. 11—SECOND-ORDER VCVS low-pass filter circuit normalized to 1 kHz by the selection of component values.

offset is not a problem, resistor $R3$ can be replaced with a short circuit between the operational amplifier output and the inverting input. If passband gain for the second-order, low-pass filter is required, then resistors $R3$ and $R4$ are used.

The second-order VCVS filter is by far the most commonly used type. Its -40-dB/decade roll-off, coupled with a high degree of stability, results in a generally good trade-off between performance and complexity.

The general form of the second-order filter transfer equation is similar to the expression for the first-order filter:

$$A_{dB} = 20\text{LOG}(A_v) - 20\text{LOG}[(\omega_o)^4 + (a^2 - 2)(\omega_o)^2 + 1]^{1/2}$$

where a is the damping factor of the circuit, and other terms are as defined earlier for the first-order case.

The damping factor (a) is determined by the form of filter circuit. For the Butterworth design the value of a is 1.414.

The passband gain for this circuit is the normal gain for any non-inverting follower/amplifier. If the output is strapped directly to the inverting input, or if $R3$ (but not $R4$) is used in the feedback network, then the gain is unity ($A_v = +1$). For

gains greater than unity ($A_v > 1$), the following is true:

$$A_v = (R3/R4) + 1$$

The cut-off frequency (f_c) is the frequency at which the voltage gain drops -3 dB from the passband gain. This gain is found from:

$$A_v = 1/[2\pi(R1R2C1C2)]^{1/2}$$

The gain magnitude (V_o/V_{in}) is found in a manner similar to the first-order case:

$$V_o/V_{in} = A_v/[1 + (f/f_c)^4]^{1/2}$$

There is no requirement in VCVS filters that like components (R or C) in the frequency selective network be made equal, but such a step simplifies the design procedure. If $R1 = R2 = R$, and $C1 = C2 = C$, then:

$$f_c = 1/2\pi RC$$

A constraint on this simplification is that the Butterworth response is guaranteed only if A_v is equal to or greater than 1.586.

Design procedure

1. Select the -3-dB cut-off frequency (f_c) based on the circuit requirements and applications.
2. Select a standard value capacitance.
3. Calculate the required resistance from:

$$R1 = 1/2\pi f_c C$$

4. Select the passband gain for f less than f_c .
5. Select a value for resistor $R4$, and
6. Calculate $R3$ from:

$$R3 = R4(A_v - 1)$$

The normalized 1-kHz trial values for doing scaling design of the second-order low-pass filter are shown in Fig. 11. The design here is based on a more complex arrangement whereby $C2 = 2C1$. Some designers maintain that this is the superior design. The same scaling rule is applied to the second-order filter as used in the first-order design.

Now that you know a little more about the low-pass filter we are going to change gears and look at the high-pass and band-pass filter circuits.

High-pass filters

The function of the high-pass filter is the inverse to that of the low-pass filter, so one can reasonably expect its frequency-response characteristic to mirror that of the low-pass filter response. Figure 12 shows the high-pass filter response with roll-off slopes of -20-, -40-, and -60-dB/decade. Compare the frequency-response curves in Fig. 12 to those in Fig. 3. The passband of the high-pass filter are all frequencies above the cut-off frequency f_c . As in the low-pass case, f_c is the frequency at which passband gain drops -3 dB; that is $A_{vc} = 0.707A_v$.

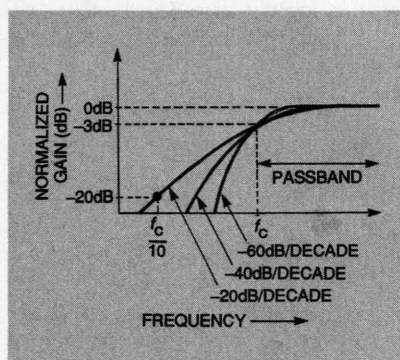


FIG. 12—FREQUENCY RESPONSE characteristic for first-order, second-order and third-order VCVS high-pass AC filters.

The cut-off frequency phase shift in a high-pass filter has the same magnitude as the low-pass case, but the sign is opposite. At f_c , the high-pass filter exhibits a phase shift of +45 degrees per 20-dB/decade of roll-off. Put another way, the phase shift is $(n \times 45\text{-degrees})$, where n is the order of the filter.

The high-pass versions of the VCVS filters are of the same form as the low-pass filter. In the case of the high-pass filter, however, impedances Z1 through Z3 are capacitances, while Z4 through Z6 are resistances (Fig. 8). In the high-pass filter the roles of the resistors and capacitors are reversed.

First-order high-pass filters

The circuit for a first-order high-pass filter is shown in Fig. 13. This circuit is identical to the first-order low-pass filter in which the roles of C1 and R1 are

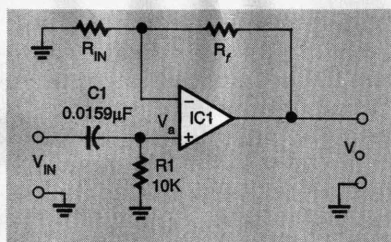


FIG. 13—FIRST-ORDER VCVS high-pass filter circuit normalized to 1-kHz by the selection of component values.

interchanged (see Fig. 10). The high-pass filter shown here is normalized for 1 kHz. Passband gain of this circuit is:

$$A_v = (R_f/R_{in}) + 1$$

The voltage at the non-inverting input of the operational amplifier (V_a) is developed across resistor R1, and is given by:

$$V_a = j2\pi f R1 C1 V_{in} / (1 + j2\pi f R1 C1)$$

The transfer equation for the circuit is:

$$V_o = A_v V_a$$

$$V_o = [(R_f/R_{in}) + 1] \times [j2\pi f R1 C1 V_{in} / (1 + j2\pi f R1 C1)]$$

And, in the traditional form, the equation becomes:

$$V_o/V_{in} = A_v j(f/f_c) / [1 + (f/f_c)]$$

As in the previous cases, the cut-off frequency f_c is found from:

$$f_c = 1/2\pi R1 C1$$

The gain magnitude of this circuit is the absolute value of the traditional form of the transfer equation:

$$V_o/V_{in} = A_v (f/f_c) / [1 + (f/f_c)^2]^{1/2}$$

The VCVS high-pass filter shown in Fig. 6 is normalized to 1 kHz. The same scaling technique is used for this circuit as was used for the low-pass filters discussed earlier.

Second-order high-pass filter

The second-order high-pass filter offers a roll-off slope of -40 dB/decade as illustrated in Fig. 12. This VCVS filter circuit (Fig. 14) is, like its low-pass counterpart, probably the most commonly used form of high-pass filter. The circuit is similar to the low-pass design except for a reversal of the roles of capaci-

tors and resistors. The cut-off frequency is the frequency at which gain falls off -3 dB, and is found from:

$$f_c = 1/2\pi [R1 R2 C1 C2]^{1/2}$$

or, in the case where $R1 = R2 = R$, and $C1 = C2 = C$:

$$f_c = 1/2\pi RC$$

The gain magnitude of the circuit is found from:

$$V_o/V_{in} = A_v [1 + (f_c/f)^4]^{1/2}$$

Example

Calculate the cut-off frequency of a filter shown in Fig. 14 in which $C1 = C2 = 0.0056 \mu F$ and $R1 = R2 = 22,000$ ohms.

$$f_c = 1/2\pi RC$$

$$f_c = 1/(7.74) \times 10^{-4} = 1,293 \text{ Hz}$$

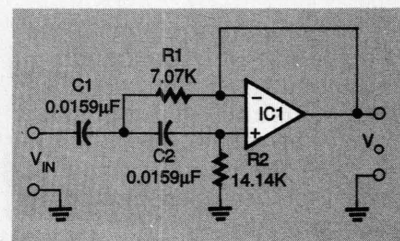


FIG. 14—SECOND-ORDER VCVS high-pass filter circuit normalized to 1-kHz by the selection of component values.

Band-pass filter

The band-pass filter is a circuit that has a passband between an upper limit and a lower limit. Frequencies above and below these limits are in the stopband. There are two basic forms of band-pass filters: Wide band pass and narrow band pass. These two types are sufficiently different that they offer different frequency response characteristics. The wide band-pass filter may have a passband that is wide enough to be called a band-pass amplifier rather than a filter and its frequency

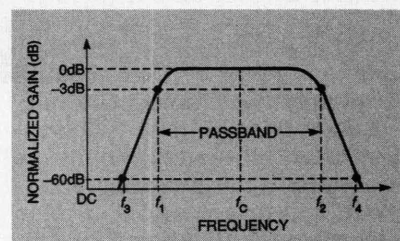


FIG. 15—TYPICAL BANDPASS curve for a wide-band filter.

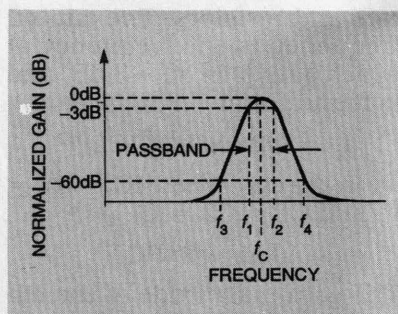


FIG. 16—TYPICAL BANDPASS curve for a narrow-band filter.

response curve is shown in Fig. 15. The narrow band-pass frequency response is shown in Fig. 16. The passband is defined as the frequency difference between the upper -3-dB point (f_2), and the lower -3-dB point (f_1). The bandwidth (BW) is:

$$BW = f_2 - f_1$$

The center frequency f_c (not to be confused with the cut-off frequency f_c) of the band-pass filter is usually symmetrically placed between f_1 and f_2 , or $(f_2 - f_1)/2$. If the filter is a very wide-band type, however, the center frequency is:

$$f_c = [f_1 f_2]^{1/2}$$

Band-pass filters are sometimes characterized by the figure of merit, or Q that is a factor that describes the sharpness of the filter, and is computed from:

$$Q = f_c/BW = f_c/(f_2 - f_1)$$

The Q of the filter tells us something of the passband characteristic. Wide-band filters generally have a Q less than 10, while narrow-band filters have a Q greater than 10.

The *shape factor* of the filter characterizes the slope of the roll-off curve, so is obviously related to the order of the filter. The shape factor (SF) is defined as the ratio of the -60-dB bandwidth to the -3-dB bandwidth:

$$SF = BW_{-60\text{dB}}/BW_{-3\text{dB}}$$

First-order band-pass filters

A wide-band first-order band-pass filter frequency response is obtained by cascading first-order, high-pass and low-pass filter circuits, as shown in Fig. 17. This arrangement overlays, or superimposes, the frequency

response characteristics of both filter stages into one stage. Figure 18 shows the filter's frequency response when cascade high- and low-pass filters are used. The low-pass filter response (solid line) is from DC to the -3-dB point at f_2 . The high-pass filter response (dashed line) is from the highest possible frequency within the range of the circuit down to the -3-dB point at f_1 . The passband is the intersection, or overlay, of the two sets: high and low-pass characteristics, which falls between f_1 and f_2 .

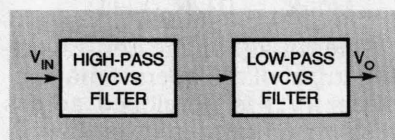


FIG. 17—BANDPASS FREQUENCY response can be achieved by cascading low-pass and high-pass AC filter stages

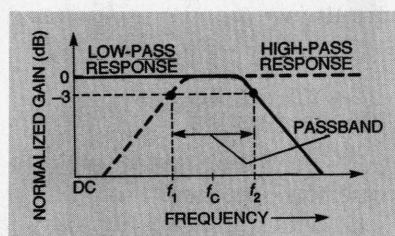


FIG. 18—THE PASSBAND is the overlapping of two separate frequency response curves.

The gain of the overall band-pass filter within the pass band is the product of the two individual gains:

$$A_{vt} = A_{VL} \times A_{VH}$$

The gain magnitude term of this form of filter is found by:

$$V_o/V_{in} = A_{vt}(f/f_1)/[(1 + (f/f_1)^2) \times (1 + (f/f_2)^2)]^{1/2}$$

where V_o is the output signal voltage; V_{in} is the input signal voltage; f is the applied frequency; f_1 is the lower -3-dB frequency; f_2 is the upper -3-dB frequency; and A_{vt} is the total cascade gain of the filter.

Cascading low and high-pass filter sections can be used to make wideband filters, but because of component tolerance and other problems it becomes less useful as Q increases above about 10 or so. For narrow-band filters a multiple-feedback-path

(MFP) filter circuit such as shown in Fig. 19 can be used. This filter circuit (Fig. 19) offers first-order performance and relatively narrow band-pass. The circuit will work for values of Q more than or equal to 10 and Q less than or equal to 20, and gains up to about 15. The center frequency of the MFP band-pass filter is:

$$f_c = (1/2\pi)[1/R_3C_1C_2(1/R_1 + 1/R_2)]^{1/2}$$

To calculate the resistor values it is necessary to first select the passband gain (A_v) and Q . It is the general practice to select values for C_1 and C_2 , and then calculate the required resistances for the specified values of f_c , A_v and Q . The resistor values are:

$$R_1 = 1/2\pi A_v C_2 f_c$$

$$R_2 = 1/2\pi f_c (2Q^2 - A_v) C_2$$

$$R_3 = 2Q/2\pi f_c C_2$$

and the gain

$$A_v = R_3/R_1(1 + C_2/C_1)$$

These equations can be simplified if the two capacitors are made equal ($C_1 = C_2 = C$), and assuming that Q is greater than $(A_v/2)^{1/2}$:

$$R_1 = Q/2\pi f_c A_v C$$

$$R_2 = Q/2\pi f_c C (2Q^2 - A_v)$$

$$R_3 = 2Q/2\pi f_c C$$

$$A_v = R_3/2R_1$$

Example

Design an MFP band-pass filter with a gain of 5 and a Q of 15 when the center frequency is 2,200-Hz. Assume that $C_1 = C_2 = 0.01\text{-}\mu\text{F}$. Solution:

$$R_1 = Q/2\pi f_c A_v C$$

$$R_1 = 217,140 \text{ ohms}$$

$$R_2 = Q/2\pi f_c (2Q^2 - A_v)$$

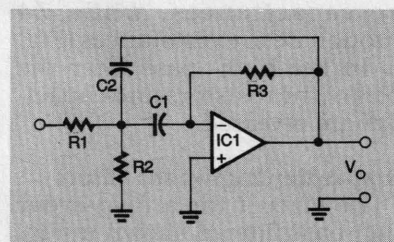


FIG. 19—MULTIPLE-FEEDBACK-PATH (MFP) filter achieves narrow-band band-pass response.

$$R2 = 15/0.062 = 240 \text{ ohms}$$

$$R3 = 2Q/2\pi f_c C$$

$$R3 = 217,140 \text{ ohms}$$

The band-pass filter is capable of being tuned using only one of the resistors. When $R2$ is varied, the center frequency will shift, but the bandwidth, Q , and gain will remain constant. To scale the circuit to a new center frequency (f_c) using only $R2$ as the change element, select a new value of $R2$ according to:

$$R2' = R2(f_c/f_c')$$

Notch filters

A band-reject or notch filter is used to pass all frequencies except a single frequency (or small band of frequencies). A common application for this circuit is to remove 60-Hz interference from sensitive electronic instruments. For example, the medical electrocardiograph (ECG) machine suffers 60-Hz line interference. These devices often include a switch-selectable 60-Hz notch filter to reject the interference.

Figure 20 shows a typical active-notch filter, while Fig. 21 shows the frequency-response curve for that circuit. Note that the gain is constant throughout the frequency spectrum except in the immediate vicinity of f_c . The depth of the notch is infi-

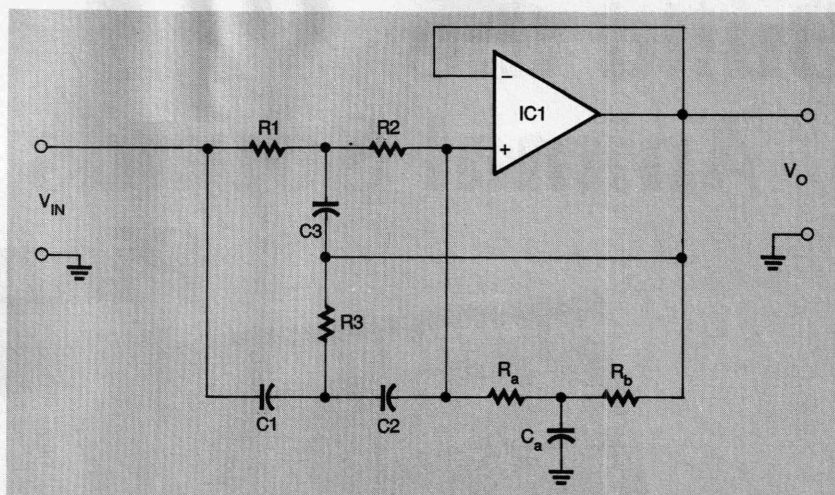


FIG. 20—A TYPICAL UNITY-GAIN, active-notch filter circuit based on the twin-tee network and an operational amplifier.

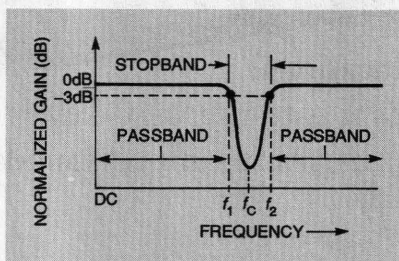


FIG. 21—FREQUENCY-RESPONSE curve for the MFP filter circuit displays a narrow frequency notch.

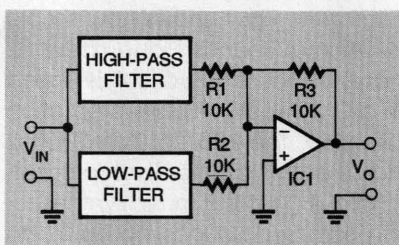


FIG. 22—BAND-REJECT FILTER is made by summing the high-pass and low-pass filter circuit elements into one circuit.

nite in theory, but in practical circuits precision matched components will offer -60 dB of suppression, while ordinary components can offer -40 to -50 dB of suppression. The resonant frequency of this notch filter is found from:

$$f_c = 1/2\pi RC$$

The gain of the circuit is unity, but the Q can be set according to the following equations:

$$Q = R_a/2R$$

or

$$Q = C/C_a$$

Bandstop filter

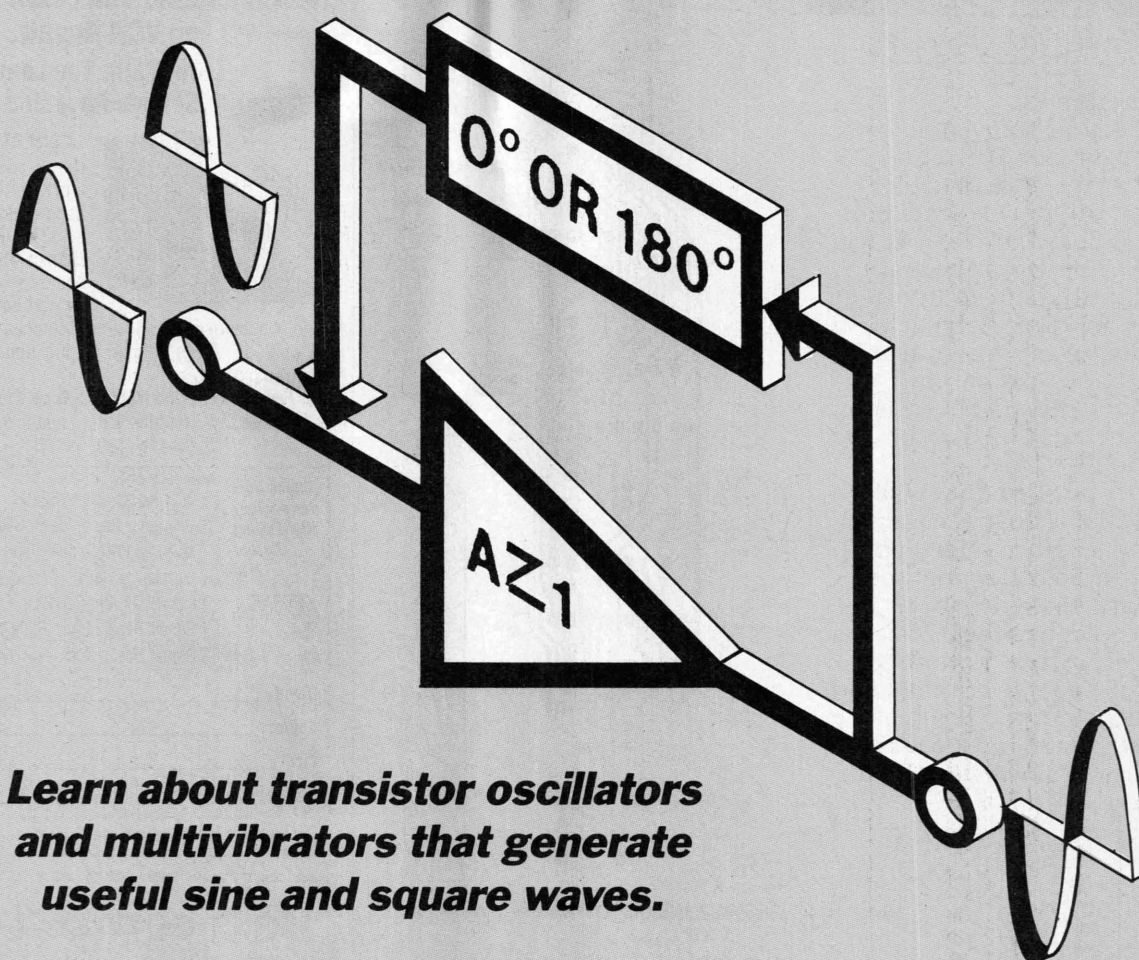
A bandstop filter is an example of a notch filter with a wider stopband. Just as the wide band-pass filter can be made by cascading high- and low-pass filters, the wide-band notch (or stopband) filter can be made by paralleling high- and low-pass filters. Figure 22 shows a bandstop filter in which the outputs of a high-pass filter and a low-pass filter are summed together in a two-input, unity-gain, inverting-follower amplifier circuit. The frequency-response curves of the two filter sections are superimposed to eliminate the undesired band. When designing the bandstop filter, select the -3-dB point of the high-pass filter equal to the upper end of the stopband, and the -3-dB point of the low-pass filter equal to the lower end of the stopband.

Summary

The subject of AC filter circuits is often presented as a mystical art. Indeed, the actual mathematical basis for these circuits is a bit esoteric; nonetheless, the truth is that most people can design and build practical filter circuits using only simple algebra. In fact, if you use the method of the normalized model even that math is avoided. Now, there is no reason to avoid using these very useful circuits in your own designs. Why not give it a try at your next opportunity.

FilterMaker Software

Now available on the *Electronics Now* FTP site is FilterMaker, a software program that is an extension of this article. FilterMaker has two subsections: Passive L-C Filters and Active Op-Amp Filters. The latter offers circuit diagrams of the filter circuits covered in this article. The component values can be changed automatically for different frequencies at the touch of a screen button. Designing the circuits leaves the computations for your computer to resolve in the blink of an eye. You can get FilterMaker by connecting to the FTP site (<ftp:germs-back.com/pub/EN>) and downloading the file FILTERS.ZIP. After unzipping, you can install the program on computers running Windows 3.0 or better. From the Program Manager, click on FILE, then RUN, type A:SETUP and press ENTER. (This assumes that the unzipped version of FILTERS.ZIP is allocated on a disk in your A drive.) An icon will appear on the Program Manager screen that you can click to activate the program.



Learn about transistor oscillators and multivibrators that generate useful sine and square waves.

OSCILLATORS

RAY MARSTON

OSCILLATORS BASED ON THE bipolar junction transistor (BJT) are the subjects of this article. Previous subjects in this series have included an article on the characteristics of the bipolar junction transistor, one on the common-collector amplifier, a third on the common-emitter and common-base voltage amplifiers (September, October, and November).

Oscillator fundamentals

An oscillator is a circuit that is capable of a sustained AC output signal obtained by converting input energy. Oscillators can be designed to generate a variety of signal waveforms, and they are convenient sources of sinusoidal AC signals for testing, control, and frequency con-

version. Oscillators can also generate square waves, ramps, or pulses for switching, signaling, and control.

Simple oscillators produce sinewaves, but another form, the multivibrator, produces square or sawtooth waves. These circuits were developed with vacuum-tubes, but have since been converted to transistor oscillators. Figure 1 is a simple block diagram showing an amplifier and a block representing the many oscillator phase-shift methods.

Regardless of its amplifier, an oscillator must meet the two *Barkhausen conditions* for oscillation: 1. The loop gain must

be slightly greater than unity. 2. The loop phase shift must be 0° or 360° .

To meet these conditions the oscillator circuit must include some form of amplifier, and a portion of its output must be fed back regeneratively to the input. In other words, the feedback voltage must be positive so it is in phase with the original excitation voltage at the input. Moreover, the feedback must be sufficient to overcome the losses in the input circuit (gain equal to or greater than unity).

If the gain of the amplifier is less than unity, the circuit will not oscillate, and if it is significantly greater than unity, the circuit will be over-driven and produce distorted (non-sinusoidal) waveforms.

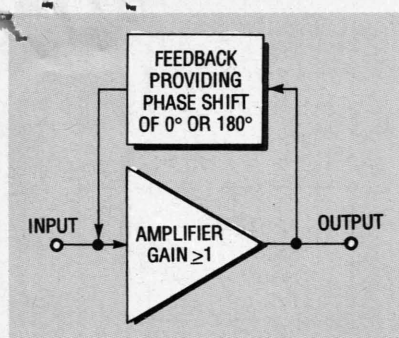


FIG. 1—OSCILLATION CONDITIONS of a sinewave generator. Loop phase shift must be 360° , and there must be adequate loop gain.

As you will learn, the typical amplifier—vacuum tube, bipolar junction transistor, or field-effect transistor—imparts a 180° phase shift in the input signal, and the resistive-capacitive (RC) feedback loop imparts the additional 180° so that the signal is returned in phase. Energy coupled back to the input by inductive methods can, however, be returned with zero phase shift with respect to the input.

Specialized oscillators such as the Gunn diodes and klystron tubes oscillate because of negative resistance effects, but the basic oscillator principles apply here as well.

RC oscillators.

Figure 2 is the schematic for a *phase-shift oscillator*, a basic resistive-capacitive oscillator. Transistor Q1 is configured as a common-emitter amplifier, and its output (collector) signal is fed back to its input (base) through a three-stage RC ladder network, which includes R5 and C1, R2 and C2, and R3 and C3.

Each of the three RC stages in this ladder introduces a 60° phase shift between its input and output terminals so the sum of those three phase shifts provides the overall 180° required for oscillation. The phase shift per stage depends on both the frequency of the input signal and the values of the resistors and capacitors in the network.

The values of the three RC ladder network capacitors C1, C2, and C3 are equal as are the val-

ues of the three resistors R5, R2, and R3. With the component values shown in Fig. 2, The 180° phase shift occurs at about $1/14 RC$ or 700 Hz. Because the transistor shifts the phase of the incoming signal 180° , the circuit also oscillates at about 700 Hz.

At the oscillation frequency, the three-stage ladder network has an attenuation factor of about 29. The gain of the transistor can be adjusted with trimmer potentiometer R6 in

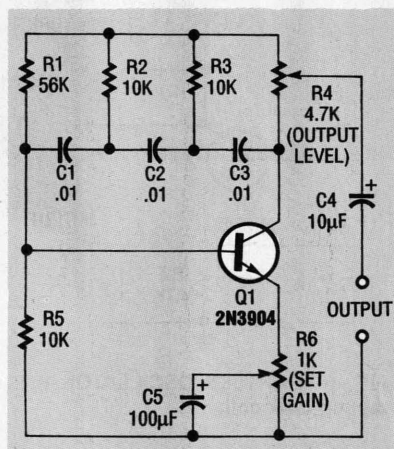


FIG. 2—PHASE-SHIFT OSCILLATOR has a three-stage RC network phase shifter.

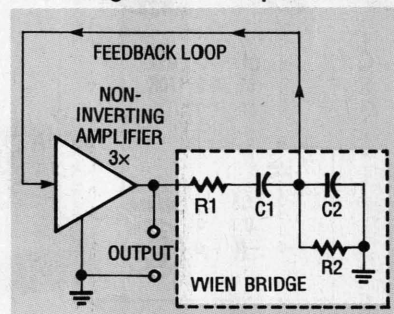


FIG. 3—WIEN BRIDGE will function as a 180° feedback network in an oscillator.

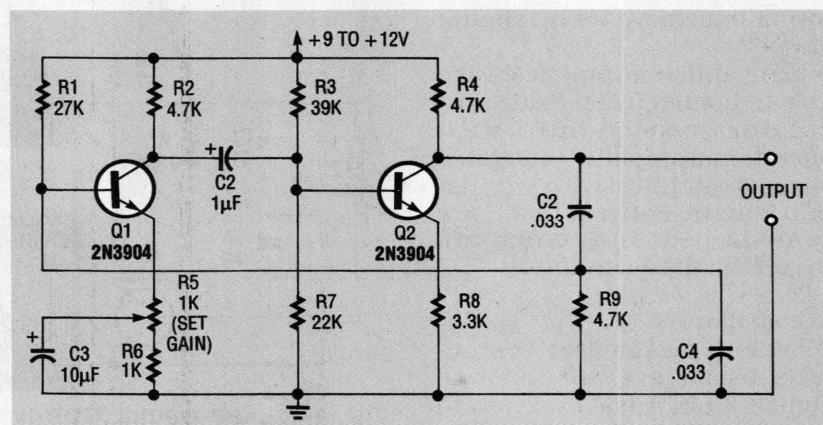


FIG. 4—WIEN-BRIDGE OSCILLATOR has an Wien-bridge network as its phase shifter.

the emitter circuit to compensate for signal loss and provide the near unity gain required for generating stable sinewaves. To ensure stable oscillation, R6 should be set to obtain a slightly distorted sinewave output.

The amplitude of the output signal can be varied with trimmer potentiometer R4. Although this simple phase-shift oscillator requires only a single transistor, it has several drawbacks: poor gain stability and limited tuning range.

There are ways to overcome the drawbacks of the phase shift-oscillator, and one of them is to include a *Wien-bridge* or network in the oscillator's feedback loop. The concept is illustrated in the Fig. 3 block diagram. A far more versatile RC oscillator than the phase-shift oscillator, its operating frequency can be varied easily.

As shown within the dotted box in Fig. 3, A Wien bridge consists of a series-connected resistor and capacitor, wired to a parallel-connected resistor and capacitor. The component values are "balanced" so that R1 equals R2 and C1 equals C2.

The Wien network is exceptionally sensitive to frequency. That shift is negative (to a maximum of -90°) at low frequencies, and positive (to a maximum of $+90^\circ$) at high frequencies. It is zero at a center frequency of $1/6.28RC$. At the center frequency, network attenuation is a factor of 3.

As a result, the Wien network will oscillate if a non-inverting, amplifier with a gain of 3 is con-

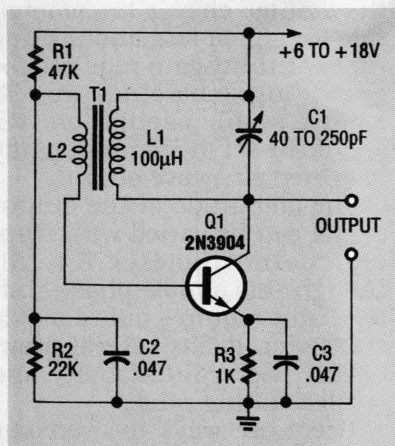


FIG. 5—TUNED-COLLECTOR oscillator has a tuned-transformer resonant tank.

nected as shown between the amplifier's output and input terminals. The output is taken between the output of the amplifier and ground.

A basic two-stage Wien-bridge oscillator schematic is shown in Fig 4. Both transistors Q1 and Q2 are configured as low-gain common-emitter amplifiers. The voltage gain of Q2 is slightly greater than unity, and it provides the 180° phase shift required for regenerative feedback. The 4.7-kilohm resistor R4, part of the Wien network, functions as the oscillator's collector load.

Transistor Q1 provides the high input impedance for the output of the Wien network. Trimmer potentiometer R5 will set the oscillator's gain over a limited range. With the component values shown, the Wien-bridge oscillator will oscillate at about 1 kHz. Trimmer R5 should be adjusted so that the sinewave output signal is just slightly distorted to achieve its maximum stability.

Many different practical variable-frequency Wien-bridge oscillators can be built with operational amplifier integrated circuits combined with an automatic gain-control feedback systems. No inductors are needed in these circuits.

LC oscillators

Resistive-capacitive sine-wave oscillators can generate signals from a few hertz up to several megahertz, but inductive-capacitive (LC) oscillators

can generate sinewave outputs from 20 or 30 kHz up to UHF frequencies.

An LC oscillator includes an LC network that provides the frequency-selective feedback between the output of the amplifier and its input terminals.

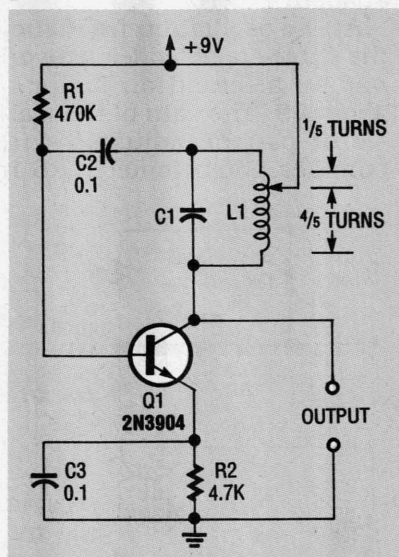


FIG. 6—HARTLEY OSCILLATOR has a tapped tank coil.

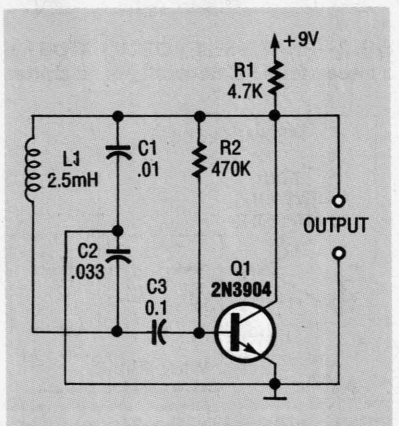


FIG. 7—COLPITTS OSCILLATOR has a dual-capacitor voltage divider.

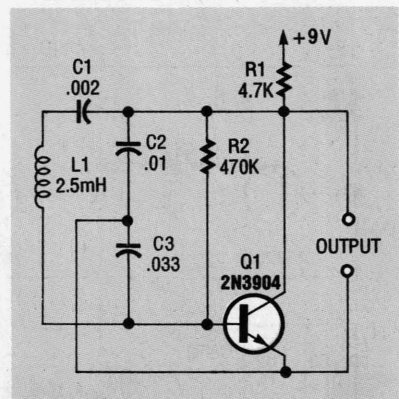


FIG. 8—CLAPP OSCILLATOR is a Colpitts oscillator with improved stability.

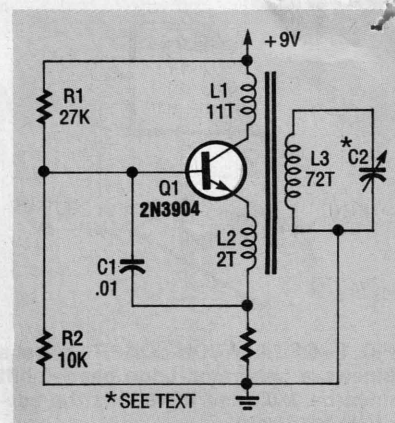


FIG. 9—REINARTZ OSCILLATOR inductively couples emitter and collector signals to its tank circuit.

Because of the inherently high *Q* or frequency selectivity of LC networks or resonant tank circuits, LC oscillators produce more precise sinewave outputs—even when the loop gain of the circuit is far greater than unity.

The tuned-collector oscillator shown in Fig. 5 is the simplest of many different LC oscillators. Transistor Q1 is configured as a common-emitter amplifier, with its base bias provided by the junction of series resistors R1 and R2. Emitter resistor R3 is decoupled from high-frequency signals by capacitor C3.

The primary turns of transformer T1 (L1) in parallel with trimmer capacitor C1 form a tuned collector resonant tank circuit. Collector-to-base feedback is provided by coil L2 in transformer T1. Coil L2, with a smaller number of turns than L1, is inductively coupled to L1 by transformer action.

The necessary zero phase shift around the feedback loop can be obtained by adjusting trimmer capacitor C1. If loop gain exceeds unity at the tuned frequency, the circuit will oscillate. Loop gain is determined by the turns ratio of L1 with respect to L2 in transformer T1.

The phase relationship between the energizing current of all LC tuned circuits and induced voltage varies over the range of -90° to +90 deg, and it is zero at a center frequency given by the formula:

$$f = 1/2\pi\sqrt{LC}.$$

Because the circuit in Fig. 5

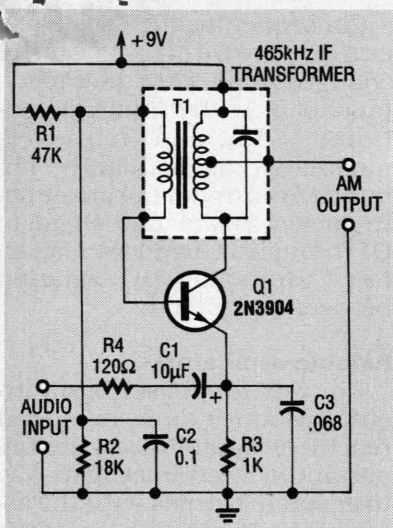


FIG. 10—BEAT-FREQUENCY oscillator that can amplitude modulate audio signals.

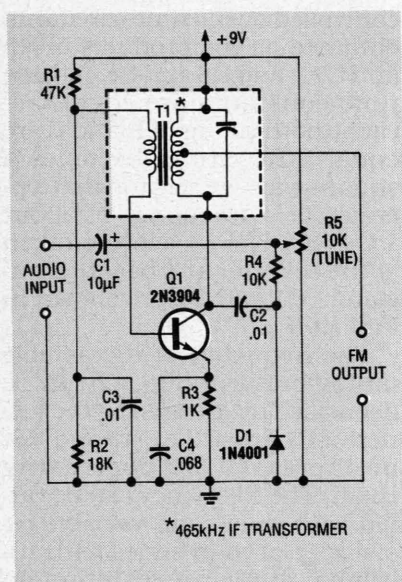


FIG. 11—BEAT-FREQUENCY oscillator with varactor tuning frequency modulates input signals.

provides a 0° overall phase shift, it oscillates at this center frequency. The frequency can be varied by trimmer capacitor C1 from 1 MHz to 2 MHz. The circuit can be enhanced to oscillate at frequencies from less than 100 Hz to UHF frequencies with a laminated iron-core transformer. The same circuit will oscillate satisfactorily in the UHF regions with an air-core transformer.

Classic LC oscillators

Figure 6 illustrates the Hartley oscillator, which is a

variation of the tuned-collector oscillator that was shown in Fig. 5. This oscillator is recognizable by the tapped coil in its tuned resonant circuit. Oscillation of the Hartley oscillator circuit depends on phase-splitting autotransformer action of the tapped coil in the tuned resonant circuit.

The tap is located on load inductor L1 about 20% of the way down from its top so that about $\frac{1}{5}$ of the turns are above the tap and $\frac{4}{5}$ are below. The positive power supply is connected to the tap to obtain the necessary

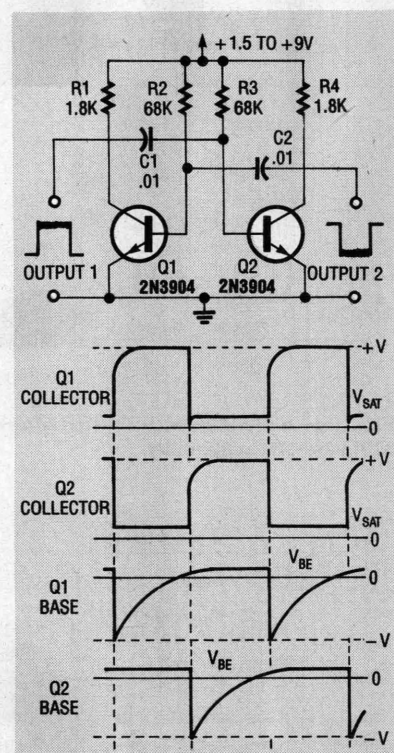


FIG. 12—ASTABLE MULTIVIBRATOR produces 1-kHz square waves (a), and waveforms at the collectors and bases of Q1 and Q2 (b).

autotransformer action.

the signal voltage across the top of L1 is 180° out-of-phase with the signal voltage across its lower end, which is connected to the collector of Q1. The signal from the top of the coil is coupled to the base (input) of Q1 through isolating capacitor C2. The oscillator will oscillate at a center frequency determined by its LC product.

The Colpitts oscillator shown in Fig. 7 is another classic cir-

cuit. It is identified by the voltage divider in its tuned resonant circuit. With the component values shown, this Colpitts circuit will oscillate at about 37 kHz.

Capacitor C1 is in parallel with the output capacitance of Q1, and C2 is in parallel with the input capacitance of Q1. Consequently, changes in Q1's capacitance (due to temperature changes or aging) can shift the oscillator frequency. This shift can be minimized for high frequency stability by selecting values of C1 and C2 that are large relative to the internal capacitances of Q1.

The Clapp oscillator, a modification of the Colpitts oscillator, shown in Fig. 8, offers higher frequency stability than the Colpitts oscillator. This is achieved by adding capacitor C1 in series with the coil in the tuned resonant tank circuit. It is selected to have a value that is small with respect to C2 and C3.

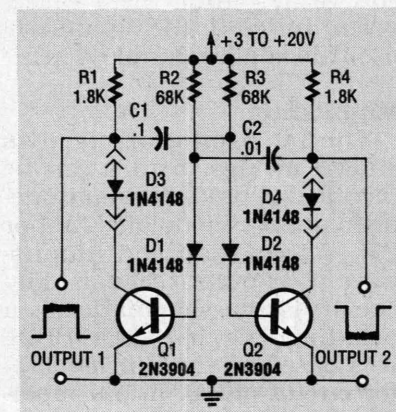


FIG. 13—ASTABLE MULTIVIBRATOR with frequency correction produces 1-kHz square waves.

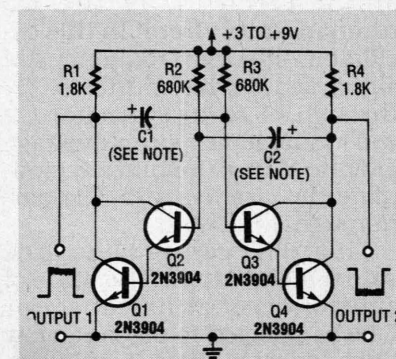


FIG. 14—ASTABLE MULTIVIBRATOR produces long-period square waves.

As a result of the presence of this capacitor, the resonant frequency of the tank and oscillator will be determined primarily by the values of L1 and C1.

Capacitor C3 essentially eliminates transistor capacitance variations as a factor in determining the Clapp oscillator's resonant frequency. With the component values shown, the Clapp oscillator oscillates at about 80 kHz.

Figure 9 shows the classic *Reinartz oscillator*. In this circuit, tuning coil L1 in the collector circuit and the tuning coil L2 in the emitter circuit are inductively coupled to tuning coil L3 in the resonant tank circuit.

Positive feedback is obtained by coupling the collector and emitter signals of the transistor through windings L1 and L2, and inductively coupling both of these coils to L3. This Reinartz oscillator oscillates at a frequency determined by L3 and trimmer capacitor C2. With the values and turns ratios given in Fig. 9, the circuit will oscillate at a few hundred kHz.

Modulation

The LC oscillator circuits shown in Figs. 5 to 9 can be modified to produce amplitude- or frequency-modulated (AM or FM) rather than continuous-wave (CW) output signals. Figure 10 is the schematic for a beat-frequency oscillator (BFO). It is based on the tuned-collector circuit of Fig. 5, but modified to become a 465-kHz amplitude-modulated (AM) BFO.

A standard 465-kHz IF transformer (T1), intended for transistor circuits, is the LC resonant tank circuit in this oscillator. An audio-frequency AM signal fed to the emitter of Q1 through blocking capacitor C2 will modulate the supply voltage of Q1 and thus amplitude-modulate the circuit's 465-kHz carrier signal.

This BFO can provide 40% signal modulation. The value of emitter-decoupling capacitor C1 was selected to present a low impedance to the 465-kHz carrier signal, while also presenting a high impedance to the low-

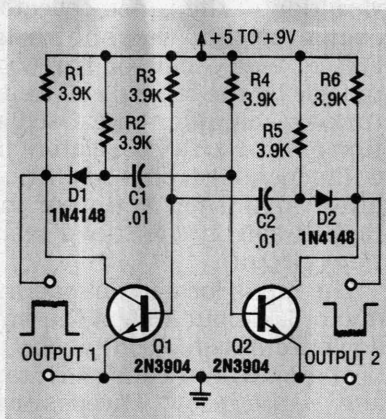


FIG. 15—ASTABLE MULTIVIBRATOR with waveform correction.

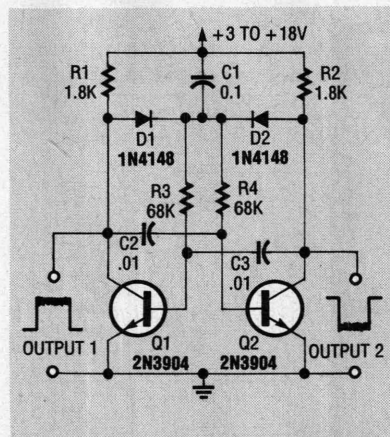


FIG. 16—ASTABLE MULTIVIBRATOR with sure-start capability.

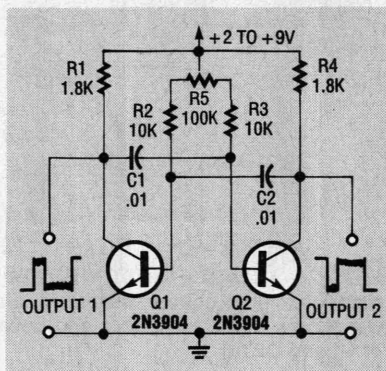


FIG. 17—VARIABLE MARK/SPACE generator runs at 1.1 kHz.

frequency modulation signal.

Figure 11 shows how the BFO circuit in Fig. 10 can be modified to become a frequency modulator. Tuning is adjusted by trimmer potentiometer R5. Silicon diode D1 functions as an inexpensive varactor diode. A 1N4001 diode frequency modulates the 465-kHz BFO circuit. Here, C2 and diode "capacitor" D1 are in series.

Consequently, the oscillator's center frequency can be changed by altering the capacitance of D1 with trimmer potentiometer R5, and frequency-modulated signals can be obtained by introducing an audio-frequency modulation signal to D1 through a C1 and R4. Capacitor C2 provides DC isolation between Q1 and D1.

Astable oscillators

Conventional oscillator circuits produce sinewaves, but repetitive square waves are important in electronics. One way to generate them is with the astable multivibrator circuit shown in Fig. 12-a.

This multivibrator is a self-oscillating regenerative switch whose on and off periods are controlled by the time constants obtained as the products of R2 and C21, and R3 and C1. If these time constants are equal (because both values of R and C are equal), the circuit becomes a square-wave generator that operates at a frequency of about $1/1.4 RC$. The waveforms taken at the collector and base of transistors Q1 and Q2 are shown in Fig. 12-b.

The frequency of the astable multivibrator in Fig. 12 can be decreased by increasing the values of C1 and C2 or R2 and R3, or increased by decreasing them. The frequency can be varied with dual-gang variable resistors placed in series with 10-kilohm limiting resistors in place of R2 and R3.

The operating frequency can, if required, be synchronized to that of a higher-frequency signal by coupling part of the external signal into the timing networks of the astable circuit. Outputs can be taken from either collector of the circuit, and the two outputs are in opposite phase. The multivibrator's operating frequency is essentially independent of power supply voltage between +1.5 and +9 volts.

The upper voltage limit is set by inherent transistor behavior: as the transistors change state at the end of each half-cycle, the base-emitter junction of one transistor is re-

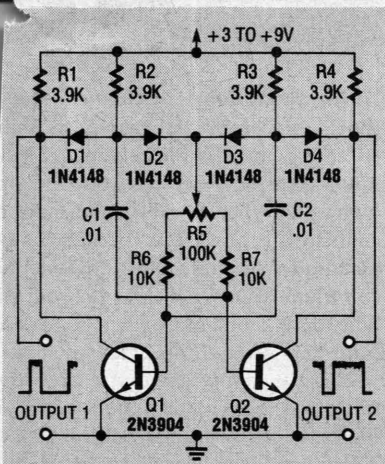


FIG. 18—VARIABLE MARK/SPACE generator features waveform correction and sure-start.

verse biased by a voltage that is about equal to the supply voltage. Consequently, if the supply voltage exceeds the reverse base-emitter breakdown voltage of the transistor, circuit timing will be affected.

This characteristic can be overcome with the circuit modifications shown in Fig. 13. A silicon diode is connected in series with the base input terminal of each transistor to raise the effective base-emitter reverse breakdown voltage of each transistor to a value greater than that of the diode.

The protected astable multivibrator will operate with any supply voltage from +3 to +20 volts. Its frequency will vary only about 2% when the supply voltage is varied from +6 to +18 volts. This variation can be further reduced to 0.5% by adding another "compensation" diode in series with the collector of each transistor, as shown in Fig. 13.

Multivibrator variations

The basic astable multivibrator shown in Fig. 12 can be modified in different ways to improve its performance or change the shape of its output waveform. Some modifications are shown in Figs. 14 to 18.

A shortcoming of the multivibrator shown in Fig. 12 is that the leading edge of each of its output waveforms is slightly rounded. The lower the values of timing resistors R2 and R3

with respect to collector load resistors R1 and R4, the more pronounced will be this waveform rounding.

Conversely, the larger the values of R2 and R3 with respect to R3 and R4, the sharper the waveform edge will be. The maximum permissible values of R1 and R2 are, however, limited by the current gains of the transistors. These gains are equal to h_{FE} multiplied either by the value of resistor R1 or R4.

One way to improve the circuit waveform, of course, would be to replace transistors Q1 and Q2 with Darlington transistor pairs and then substitute timing resistance values that are as large as permissible. That is done in the long-period astable multivibrator that is shown in Fig. 14.

Resistors R2 and R3 can have any value between 10 kilohms and 12 megohms, and the multivibrator will run from any supply voltage between +3 and +18 volts. With the R2 and R3 values shown in Fig. 14, the multivibrator's total period or cycle time is about 1 second per microfarad when C1 and C2 have equal values. This multivibrator generates sharp-cornered square waves.

The square waves with the rounded leading edges produced by the multivibrator shown in Fig. 12 are caused by an inherent characteristic of the transistor. As each transistor is switched off, its collector voltage is prevented from switching abruptly to the positive supply value. This is due to the loading between that collector and the base of the adjacent conducting transistor from timing capacitor cross-coupling.

This characteristic can be altered, and sharp square waves can be obtained by effectively disconnecting the timing capacitor from the collector of its transistor as it turns off. That improvement is shown in Fig. 15, a schematic for a 1-kHz astable multivibrator. It includes diodes D1 and D2 that disconnect the timing capacitors at the moment of switching.

The important time con-

stants of the multivibrator in Fig. 15 are also determined by C1 and R4, and C2 and R1. The effective collector loads of Q1 and Q2 are equal to the parallel resistances of R1 and R2, and R5 and R6, respectively.

Basic astable multivibrator operation depends on slight differences in their transistor characteristics. Those differences cause one transistor to turn on faster than the other when power is first applied, thus triggering oscillation.

If the multivibrator's supply voltage is applied slowly by increasing it from zero, however, both transistors could turn on simultaneously. If this happens, the oscillator will be a *nonstarter*.

The possibility of nonstarting can be avoided with the "sure-start" astable multivibrator circuit shown in Fig. 16. There, the timing resistors are connected to the transistor's collectors so that only one transistor can conduct at a time, ensuring that oscillating will always begin.

All the astable multivibrator circuits shown so far are intended to produce symmetrical output waveforms, with a 1 to 1 ratio of square wave to space (1:1 mark/space ratio). Non-symmetrical waveforms can be obtained by installing one set of RC astable time constant components that is larger than the other.

Figure 17 shows a 1.1 kHz variable mark/space ratio generator. The ratio can be varied over the range 1 to 10 with trimmer potentiometer R5. However, the leading edges of the output waveforms of this circuit could be too round for some applications when mark/space control is set at its extreme position. Also, this generator could be difficult to start if the supply voltage is applied slowly to the circuit.

Both of those drawbacks can be overcome with the modifications shown in the schematic of Fig. 18, another 1.1-kHz variable mark/space ratio generator. The circuit includes both sure-start and waveform-correction diodes.

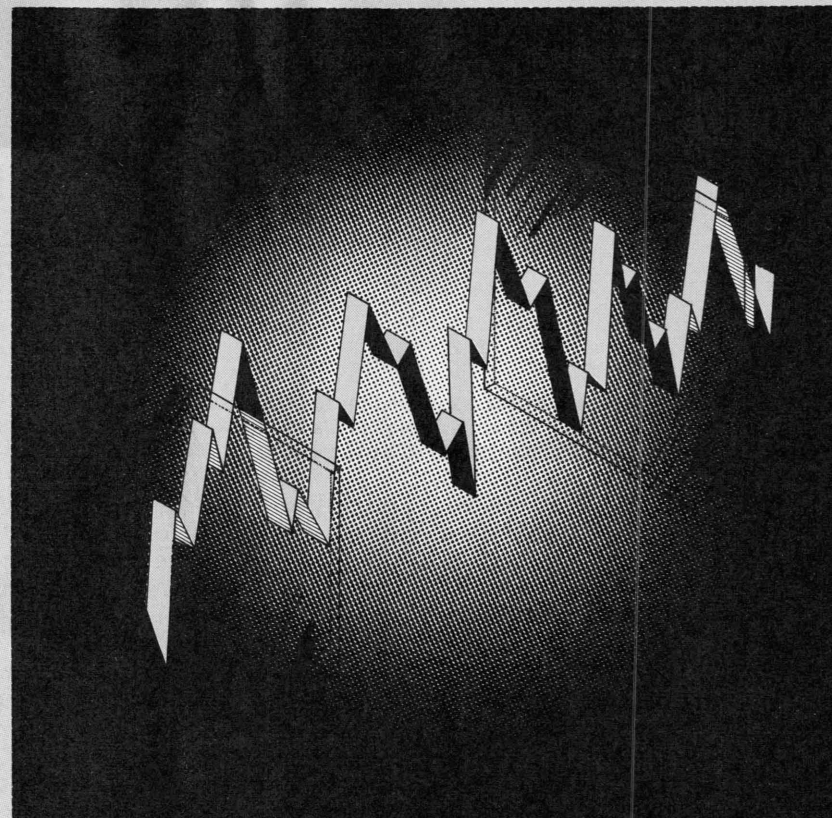
WE RECOGNIZE TWO KINDS OF ELECTRONIC noise—external and internal. External noises can be man-made—originating from such sources as motors, relays, and auto ignition—or natural—originating from lightning or aurora. Its effects can usually be eliminated by proper shielding. By contrast, internal noise is generated within your circuit; it can have many different causes and the best way to combat it is to pick your components with care.

Think of internal noise as a limiting factor on the smallest signal that can be amplified by your circuit. We can't eliminate it entirely, but we can minimize its effect. You will not have to worry too much about internal noise in your digital circuits or power amplifiers, but if you're designing or building audio amplifiers or sensitive instruments for scientific experiments, it will be very important to you. As you might expect, internal noise shows up when the input signals are small. It's still there when the input signals get bigger although it's likely to be masked.

It's most important to deal with noise in the first stage. If you deal with it there, you won't have to worry about subsequent stages. Internal noise is completely random; we can't predict what it will do to the output at any given time. Passive components such as resistors usually have a single noise mechanism, but transistors can have three or more sources. Figure 1 shows what random noise might look like on an oscilloscope.

Properties of random noise

Two properties can be determined for random noise: the average energy (or power) produced by the noise, and the average frequency distribution of this energy—commonly called the power spectrum. The average power can be thought of as the net effect of many actions in the random phenomenon. (It's analogous to the square of the rms value of a sine wave, which refers to the sine wave's



ELECTRONIC NOISE

If you know what causes internal electronic noise you can select components to minimize its effect on your circuit

average effect without considering time.)

It's impossible to predict noise amplitude or phase at any given time. Consequently, it is customary to work with a noise frequency spectrum—the average noise level as a function of frequency. Independent noise voltages do not add linearly. If two sources are independent, the noise power is added, not the voltage. Because of their random phases, the sources can be in phase and add, and

they can be out of phase and destructively interfere with each other. The average over phases is calculated by taking the root-mean-square sum of the sources. Average amplitudes are squared and added, and the total noise is the square root of the result. (This is also known as adding in quadrature.) For two equal sources, the sum has an amplitude that is twice that of the individual amplitudes.

Of the many possible power

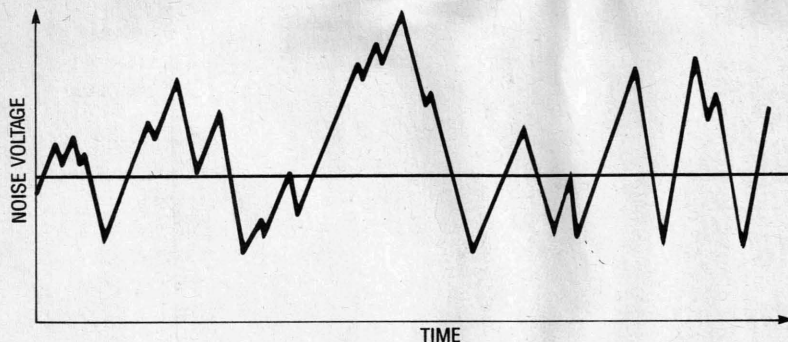


FIG. 1—RANDOM INTERNAL NOISE as viewed on an oscilloscope. The power spectrum of thermal noise is flat. Whenever noise has a flat power spectrum, it can be considered as containing components of all frequencies and is called "white" noise.

spectrums of random waves, we will be concerned with two types: a power spectrum that is "flat," that is, the power density at all frequencies is equal; and a power spectrum that varies with frequency. The most basic type of noise, thermal noise, has a flat frequency spectrum and is called "white" noise. If its spectrum varies with frequency, it's called "pink" noise.

In noise analysis, the important variable is noise power, the square of the noise voltage divided by the characteristic impedance. However, for circuit modeling, it is usually easier to work with the noise voltage, e_n . Noise power is related to the frequency bandwidth, B , so the noise voltage is proportional to the square root of the bandwidth. Noise power is measured in units of microwatts per hertz, so noise voltages are expressed in units such as nanovolts per root hertz, with the square root obtained from the power-to-voltage conversion. Because noise power varies with frequency, noise specifications must include the measurement frequency.

Figure 2 is an equivalent circuit for a signal source-amplifier combination. The signal source could be an antenna, microphone, or a sensor such as a strain gauge, thermocouple, or RTD. The input signal source is shown as a voltage generator, V_s , in series with a source noise voltage, e_{ns} , and a source impedance, Z_s . The amplifier input is modeled as an amplifier noise voltage, e_{na} , amplifier current noise, i_{na} , and amplifier in-

put impedance, Z_a . In a properly designed circuit, only the first stage noise need be considered; the gain of the first stage is high enough so that noise from succeeding stages can be neglected.

In addition to amplifying the signal, the succeeding stages will also act as filters. Because noise is broadband, it is advisable to design for the narrowest possible bandwidth to limit the output noise. It is also important to match the amplifier input impedance with the signal source output impedance to prevent signal loss and a reduction in the signal-to-noise (S/N) ratio. Thus, the input impedance of the signal source determines the choice of amplifier circuit. For example, FET's are selected where high input impedances are required and bipolar transistors are selected where that parameter is not quite as important.

Many different noise sources can be present in a single device so the output spectrum can be complex. Figure 3 maps the regions where the various noise types predominate in the frequency domain. There are many different designations for

the same type of internal noise

Thermal noise

Thermal noise is the most common type of noise in electronic components. (It is also called resistive noise, Johnson noise, and Nyquist noise). Because it is present in every electric conductor (whether or not it is connected), it is present in every electric circuit. Thermal noise is due to the random motion of free electrons within a conductor with resistance. In addition to their forward motion, electrons exhibit random (or Brownian) motion, which introduces current fluctuations. In an ideal resistor, the thermal noise voltage is given by

$$e_n = \sqrt{4kTRB}$$

Where e_n = mean value of noise voltage, k = Boltzmann's constant = 1.38×10^{-23} Joules/Kelvin (absolute), T = temperature of conductor in Kelvins, (Room temperature is approximately 300 K), R = resistance (or real part of a complex impedance) in ohms, B = frequency bandwidth in hertz = $f_2 - f_1$.

kT represents the average available thermal energy. The appearance of the term kT in most noise expressions suggests that cooling might be a suitable noise reduction technique. Unfortunately, cooling individual components is difficult and often impractical. Moreover, as will be noted in the discussion of semiconductors, some noise actually increases with cooling.

Shot noise

Shot noise is present in any electronic device where electrons move across a potential

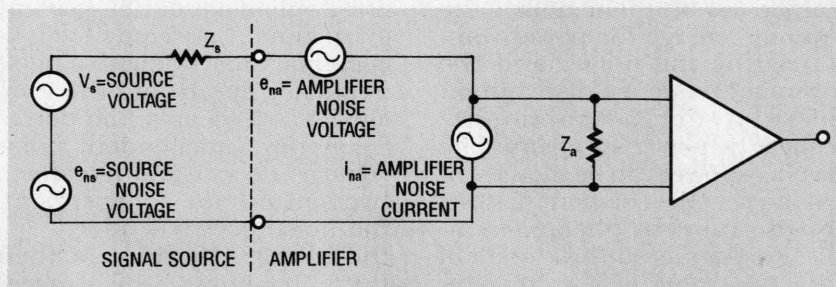


FIG. 2—NOISE EQUIVALENT CIRCUIT for a source-amplifier combination.

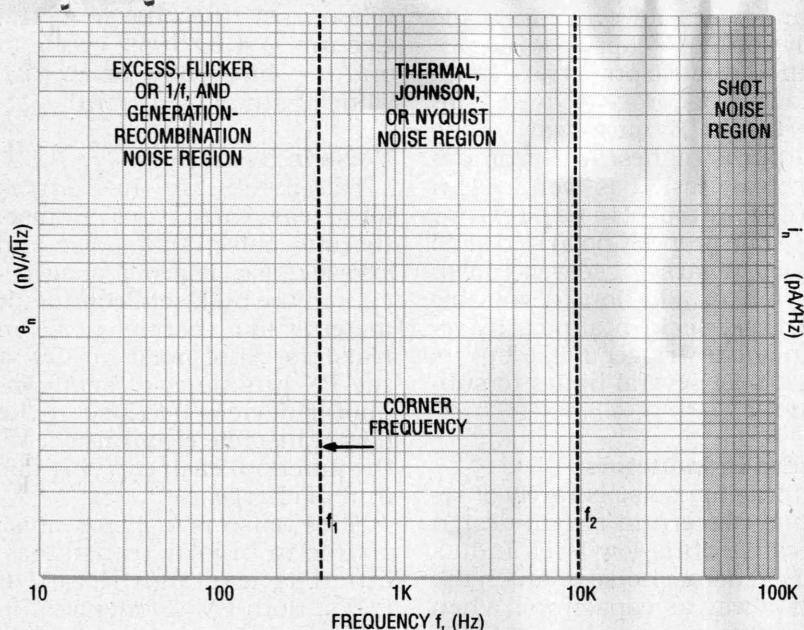


FIG. 3—MAP OF REGIONS OF INTERNAL NOISE in the frequency domain. The variable e_n is the equivalent short-circuit noise voltage and the variable i_n is the equivalent open-circuit noise current.

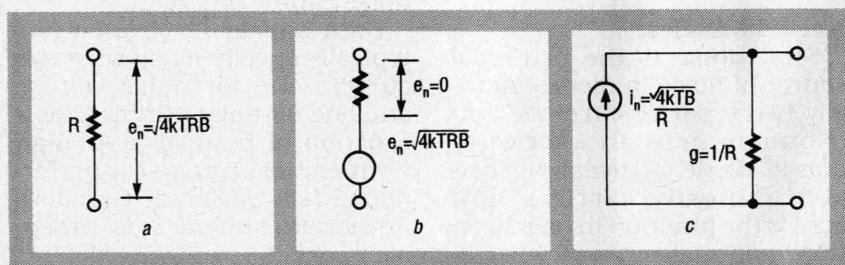


FIG. 4—NOISE VOLTAGE OF A RESISTOR and its equivalent circuit; (a) thermal noise across resistor; (b) voltage equivalent circuit for a resistor; (c) current equivalent circuit.

barrier in a random manner. For example, a noise current flows through a diode as a result of current pulses produced by individual charge carriers.

Excess noise

Another common noise source in electron devices is excess noise. (It is also known as flicker noise, $1/f$ noise, current noise, contact noise, or defect noise.) This noise is inversely related to frequency. A $1/f$ spectrum is one in which noise power amplitude increases linearly with decreasing frequency; the lower the frequency, the higher the noise level. Excess noise occurs because the resistive material is not uniformly distributed in the component. Thus some electrons encounter higher resistance than others. These non-uniform paths lead

to current flow variations or noise. Because it is electron-flow dependent, excess noise is proportional to the current through the resistor. In active devices the causes are more complex, but they relate to crystal lattice and surface effects. Excess noise is generally expressed by a noise index in decibels.

Generation-recombination

The doping centers of semiconductors also generate noise. Electron-hole pair generation and recombination at these centers creates generation-recombination (GR) noise. It is frequency-dependent and has a complex spectrum. For a single site, the spectrum is considered to be $1/f$ although this is an approximation. When the noise from a large number of sites of

varying activity levels is added, the spectrum becomes more complex and can vary with temperature in very unpredictable ways.

Two other potential noise sources are avalanche noise and crystal defects (popcorn noise). Avalanche noise occurs when a PN junction is reverse biased into breakdown, producing a current avalanche. This event can damage the junction, permanently increasing its noise level. Popcorn noise is caused by a temporary breakdown within the device. It shows up as a current flow with a duration varying from milliseconds to seconds. This noise source is difficult to characterize because it is intermittent. However, it can be observed on an oscilloscope.

Avalanche and popcorn noise will not be covered in greater detail because they can usually be avoided by proper care in device selection and handling. You can avoid this noise by operating semiconductor devices within their rated voltage and current limits.

Noise in resistors

Figure 4 illustrates noise equivalent circuits for a resistor. Figure 4-a represents thermal noise across the resistor, Fig. 4-b is the voltage equivalent circuit of the resistor shown as a voltage source in series with a noiseless resistor. Figure 4-c is the shot noise equivalent circuit of a resistor shown as a current generator in parallel with a noiseless resistor. Referring to the thermal noise equation, it should be noted that kT is roughly 4×10^{-23} Joules at room temperature, so

$$e_n = \sqrt{(R/64)B}$$

Thus a 64-ohm resistor has a noise of $1 = \text{nV}$ per root hertz. Another useful relationship is that a 1-K resistor has a noise level of about 4 nanovolts per root hertz. Thus a 20K resistor has an average noise level of 2.5 millivolts when measured over a 20-kHz (audio) bandwidth.

Excess noise can also be present in a resistor. It can be reduced by minimizing the voltage drop across the resistor.

**TABLE 1—
RESISTOR EXCESS NOISE**

Material	Excess noise (dB)
Metal oxide	-15 to -40
Wirewound	-15 to -40
Carbon-film	-5 to -25
Carbon composition	+10 to -20

Excess noise is related to resistor materials, as shown in Table 1. Proper resistor selection can help to keep it low.

Noise in capacitors

Capacitors introduce two sources of electrical noise: leakage and dielectric loss. Leakage noise occurs because capacitors, like resistors, have finite resistance values. Fortunately, that resistance is effectively in parallel with the capacitor's capacitive impedance, a much smaller value, which effectively shorts the leakage noise.

Dielectric losses occur as a capacitor charges and discharges. In each cycle a small portion of the stored energy is dissipated in the dielectric medium. This dissipation is characterized by a factor called the dielectric loss angle, which generally corresponds to the fraction of energy dissipated. The loss angle is related to the dielectric, ranging from 10^{-2} to 10^{-4} . The noise voltage for capacitors is:

$$e_n = \sqrt{4kTdB/\omega C}$$

where C = capacitance in farads, d = the dielectric loss angle, and $\omega = 2\pi f$.

Noise can be minimized by choosing a capacitor based on its dielectric loss angle. Table 2 lists typical dielectric loss angles for capacitors.

If noise is to be reduced in large-value capacitors, tantalum capacitors are preferred because tantalum oxide has a lower dielectric loss than alumi-

**TABLE 2—CAPACITOR
DIELECTRIC LOSS ANGLES**

Dielectric	Approximate loss angle (d)
Teflon	10^{-4}
Mica	10^{-3}
Paper	10^{-2}

num oxide. However, loss angle for tantalum capacitors varies with dielectric properties, manufacturing process, and temperature. By contrast, noise output is highest in aluminum electrolytics. It is well known that electrolytic capacitors must be properly polarized at all times. What's not so well known is that in addition to possible damage, accidental polarity reversal can trigger noise bursts that take several hours to subside.

Noise in inductors

Inductor noise is rarely of serious concern in circuit design because of its low level. Inductors generate noise in much the same way as capacitors. When they store energy, some of it is dissipated, creating a noise current, but it is generally at very low levels.

Noise in diodes

Shot noise is the principal source of noise in diodes. Actually two separate currents contribute to noise in a forward-biased diode: thermally generated minority carriers flow across the junction in one direction, and majority carriers dif-

Where q = the charge of the electron = 1.6×10^{-19} coulomb and I = bias point or average value of junction current.

Noise in transistors

Transistors are the sources for at least three types of noise: thermal noise, shot noise and excess noise. Thermal noise occurs in the bulk semiconductor material like thermal noise in resistors. Shot noise occurs at the PN junctions of semiconductor devices because of the flow of discrete electrons across the junction, each carrying one unit of charge.

Shot noise is a major noise source in bipolar transistors, but it is less significant in FET's. Both PN junctions in bipolar transistors conduct significant current, but because FET's do not have current flowing across PN junctions, shot noise can be ignored.

Data sheets for transistors typically specify noise with two curves, one for noise voltage and one for noise current, as a function of frequency. At high frequencies, transistor performance falls off, so the signal-to-noise ratio suffers. It is advisable to choose transistor devices that operate in their mid-band range, although for very high frequency circuits this may not be possible.

Bipolar transistors

Figure 5 is a simplified hybrid pi equivalent circuit of a bipolar transistor. The generator $g_m V_b$ produces a current proportional to the voltage across resistor r_{be} . In more familiar terms, $\beta = g_m r_{be}$. For small signals, $g_s m$ can be shown as qI_C/kT .

The resistors represent internal resistances in the transistor. Each has an associated thermal noise generator. Resistor r_{in} represents the base contact resistance, typically a few hundred ohms. Because it occurs at the input, its noise contributes directly to the transistor noise. Resistor r_{be} is usually very large but, because it is in parallel with capacitor C_{bc} , the noise is shunted, and can be neglected. Similarly, r_{ce} is large

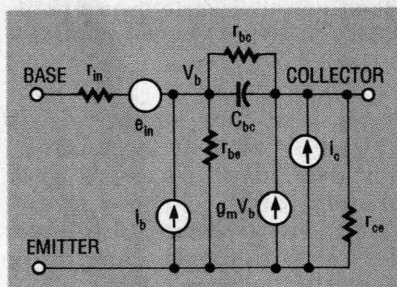


FIG. 5—HYBRID PI EQUIVALENT circuit of a transistor.

fuse across the barrier in the opposite direction. The size of those currents depends on the voltage. With no applied voltage, the two currents are opposite and equal, and the average net current is zero. However, the two currents are independent, and each contributes noise, even when their sum is zero. In a reverse-biased diode only leakage current is a factor in noise.

The noise current from shot noise in a diode is given by

$$i_n = \sqrt{2qIB}$$

and in parallel with the output, so its noise can be neglected. The effective base-emitter resistance, r_{be} , is $\beta kT/qI_c$. Because it is not a real resistance, it does not contribute noise.

The total transistor noise can be represented by a voltage generator and two current sources at the transistor input and output. The input voltage generator represents the noise from resistor r_{in} . The input current generator accounts for the shot current across the base-emitter junction as well as the generation-recombination noise. The output current generator accounts for the shot noise across the base-collector junction.

Generation-recombination noise in a transistor is generally represented as a single spectrum source added to the base-collector shot noise. It is given by

$$I_{gr} = \sqrt{(K I_b / f) B}$$

Where K = a constant for the transistor and I_b = the base current.

Generation-recombination activity is proportional to current. Because the recombination sites are dispersed throughout the transistor, rigorous determination of an equivalent voltage can be complicated. The base resistance, r_{in} is often used for this; mismatches are absorbed in the constant K . High $1/f$ noise often indicates faults in the device and is correlated with low transistor reliability.

Typical noise curves for a bipolar transistor are shown in Fig. 6. As the collector current increases, transistor gain rises, so the noise voltage drops. However, the noise current (shot noise) rises with the increased current. The relative sizes of these two sources must be considered in choosing the minimum noise operating point. At low frequencies, $1/f$ or flicker noise dominates. Where graphs are not given, a third parameter, the corner frequency, might be given. This is the frequency at which $1/f$ is equal to the mid-range noise.

Table 3 gives examples of mid-range voltage and current

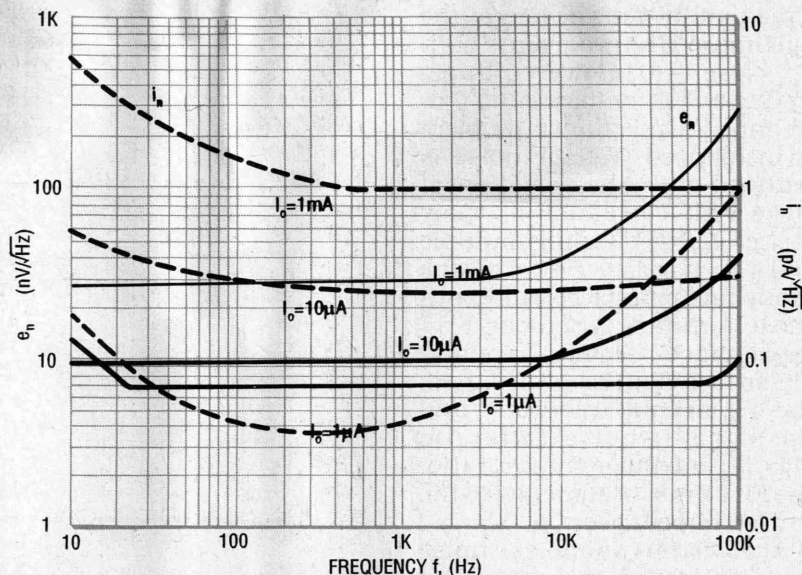


FIG. 6—TYPICAL NOISE CURVES FOR A BIPOLAR TRANSISTOR.

noises for typical bipolar transistor. At a lower collector current, for example 1 microampere, the 2N4250 noise voltage rises to 2.2 nanovolts per root hertz, while the noise current drops to 0.5 picoampere per root hertz. Because the relative contributions to total noise from noise voltage and noise current depend on the input source impedance, the optimal operating current also depends on this impedance.

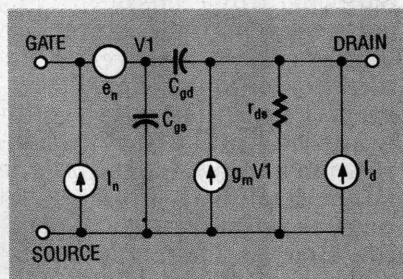


FIG. 7—NOISE EQUIVALENT CIRCUIT for a JFET.

TABLE 3—TYPICAL BIPOLAR TRANSISTOR NOISE VALUES

Transistor	I_C	Noise voltage	Noise current
2N404	1 mA	2.0 nV/ $\sqrt{\text{Hz}}$	1.8 pA/ $\sqrt{\text{Hz}}$
2N4250	1 mA	1.6 nV/ $\sqrt{\text{Hz}}$	1.2 pA/ $\sqrt{\text{Hz}}$
2N4124	1 mA	1.8 nV/ $\sqrt{\text{Hz}}$	1.4 pA/ $\sqrt{\text{Hz}}$

Field-effect transistors

Field-effect transistors (FET) have noise sources that are similar to those in bipolar transistors, but they have different levels. Figure 7 is an equivalent circuit of a JFET for noise analysis. The current generator $g_m V_1$ amplifies the signal and produces a current proportional to the voltage V_{gs} . The value r_{ds} represents the channel resistance, C_{gd} and C_{gs} represent the capacitance of the junction, and C_{ds} represents the distributed channel capacitance.

Most of the voltage noise in an FET is caused by the bulk resistance of the channel. Because this channel resistance is distributed over the entire channel, this noise is fed back to the input via the transconductance, g_m , in a complex manner. A detailed analysis leads to the relationship

$$e_n = \sqrt{4kT (0.7/g_m) B}$$

Where $0.7/g_m$ is the noise coefficient for a FET.

In most cases, this formula provides satisfactory estimates of the voltage noise. However, at high frequencies, the channel noise is also transmitted through C_{gd} . This shows up as an apparent increase in current noise at high frequencies.

To minimize e_n , you can maximize g_m by choosing a FET with a large g_m and biasing it to

a maximum. This occurs at the maximum drain current, with V_{gs} close to 0. However, in selecting a FET for maximum g_m , an input-capacitance problem is introduced. Voltage noise is reduced with the selection of large FET's because of their higher g_m , or by using multiple FETs in parallel. However, the input capacitance rises in both those options. When the FET capacitance becomes comparable to the source impedance, the signal coupling drops, and the signal-to-noise (S/N) ratio falls. To maximize the S/N ratio, the FET must be matched to the source impedance.

At lower frequencies, noise from generation-recombination centers adds to the noise voltage. As in bipolar transistors, the $1/f$ noise amplitude cannot be calculated so it must be measured. It has a complex temperature dependence because majority and minority carriers have contributions that vary with temperature. Because of the temperature dependence, moderate cooling of an FET can lead to an increase in noise.

The main source of current noise in an FET is the gate leakage current that appears directly at the gate. This is given by the equation:

$$i_n = \sqrt{2qI_g B}$$

As with bipolar transistors, FET noise is specified by graphs of noise current and noise voltage as a function of frequency. A typical characteristic is shown in Fig. 8.

Single numbers can be given for midrange noise voltage and noise current. Because $1/f$ noise is a problem only at low kilohertz frequencies, and reflected noise is generally quite small, these two numbers are usually usable over a wide range of frequencies. Because of their low junction current, JFET's emit much lower noise current than bipolar transistors while JFET noise voltage is typically slightly higher. Table 4 gives some typical noise figures for JFET's.

The noise equivalent circuits for MOSFET's are similar to those for JFET's. However, a MOSFET does not have a PN

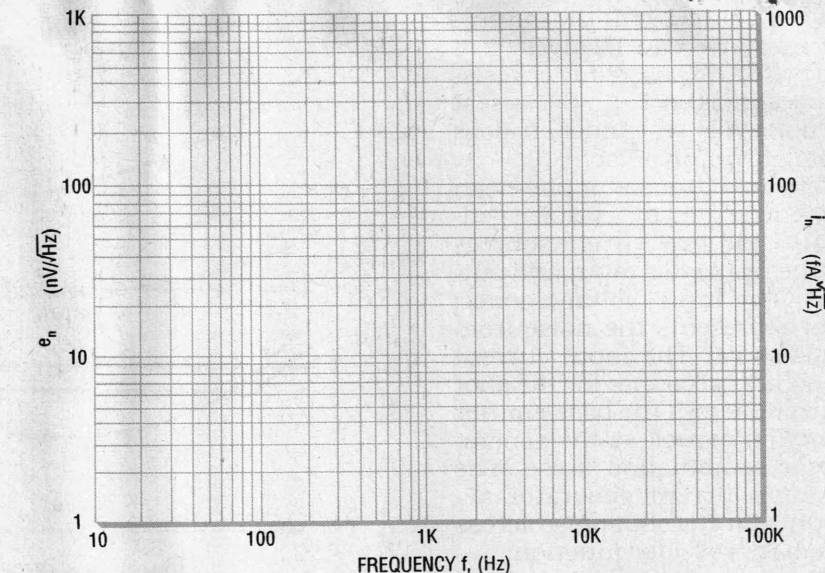


FIG. 8—TYPICAL NOISE CURVES FOR A JFET.

junction so it does not exhibit leakage current or shot noise. However, MOSFET's have a much larger $1/f$ noise voltage, 10 to 1000 times that of a JFET. For this reason, MOSFET's are used only in applications requiring very high source impedances and frequencies.

Noise in IC's

Noise equivalent circuits for integrated circuits consist of equivalent circuits for their included components. IC's exhibit slightly higher noise levels than equivalent discrete device circuits because their fabrication requires tradeoffs that err in the direction of higher noise. However, these differences are diminishing because of IC fabrication improvements, and the slight handicap in terms of noise is offset by the IC's benefits such as lower power consumption, higher reliability, and smaller size.

The noise generating ele-

TABLE 5—TYPICAL ANALOG IC NOISE VALUES

Device	Noise voltage	Noise current
741 op amp	25 nV/ $\sqrt{\text{Hz}}$	0.7 pA/ $\sqrt{\text{Hz}}$
OP-37 op amp	3 nV/ $\sqrt{\text{Hz}}$	0.4 pA/ $\sqrt{\text{Hz}}$
LF357 op amp	12 nV/ $\sqrt{\text{Hz}}$	10 fA/ $\sqrt{\text{Hz}}$
LFET input	$\sqrt{\text{Hz}}$	$\sqrt{\text{Hz}}$

ments in IC's are concentrated at the input stage, so noise modelling should focus on the input device, whether it is a JFET or a bipolar transistor. For simplicity, noise specifications of IC's are lumped together as an e_n and i_n referred to the input. The noise current might not be given for a JFET input IC because it is so small. Table 5 gives typical values for some popular IC's. The corner frequency, below which $1/f$ noise is large, is only 2.7 Hz.

Designing for low noise

Ideally, noise should be one of the first considerations in designing a circuit. Many noise-reduction techniques are fundamental to the design process. Most noise sources are wide-band so the circuit should have the narrowest possible band-pass. Frequency shifters can reduce the bandwidth. By converting RF to a lower-frequency

TABLE 4—TYPICAL JFET NOISE VALUES

JFET	Noise voltage	Noise current
2N4416	2.0 nV/ $\sqrt{\text{Hz}}$	2.0 fA/ $\sqrt{\text{Hz}}$
2N4469	2.7 nV/ $\sqrt{\text{Hz}}$	7.0 fA/ $\sqrt{\text{Hz}}$
2N5266	12 nV/ $\sqrt{\text{Hz}}$	5.0 pA/ $\sqrt{\text{Hz}}$

F, it is easier to select only the desired frequency range. At very low frequencies, where $1/f$ noise is likely to cause problems, it might be desirable to chop or upconvert that low-frequency signal to a higher band, especially if the signal source can be modulated.

A second major consideration is the selection of the first-stage active device, which usually depends on the frequency that the device must operate at and desired input impedance. For simple projects, the obvious choices are bipolar transistors or JFET's. For special projects such laboratory equipment, the designer might wish to use more exotic (and expensive) devices such as Josephson junctions, superconducting quantum interference devices (SQUID's), or tunnel-diode parametric amplifiers, all of which offer very low-noise inputs.

The crucial factor in bipolar transistor or FET selection is input impedance; FET's are inherently high-impedance devices, while bipolar transistors are lower impedance devices. Bipolars are favored for impedances in the 100-ohm to 1 megohm range, JFET's are satisfactory above 1K, while MOSFET's are the best choices for impedances above 1 megohm. If the input conditions are not established or subject to variation, as in an oscilloscope input, a JFET is preferable. JFET's offer much lower noise current than bipolar transistors, but their noise voltage is only slightly lower. Except when they must operate at very high frequencies, MOSFET's are at a disadvantage because of their large $1/f$ noise.

Once a device type is selected, the next step is to choose a specific device. This can be done by looking at the noise specifications on the data sheets. For best noise performance, it is advisable to choose a device that is operating in its midrange. Also, high-gain devices generally exhibit lower noise.

It is important that signal impedance be matched. Compare the source impedance (a func-

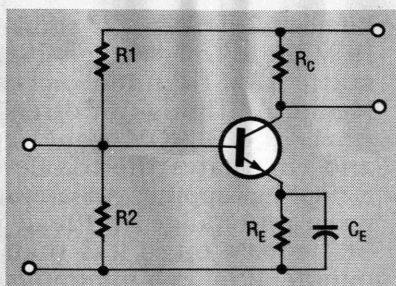


FIG. 9—COMMON-EMITTER AMPLIFIER with traditional biasing.

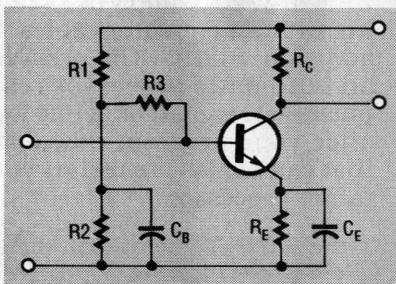


FIG. 10—COMMON-EMITTER amplifier with low-noise biasing. Noise from resistors R1 and R2 is shunted to ground by the capacitor.

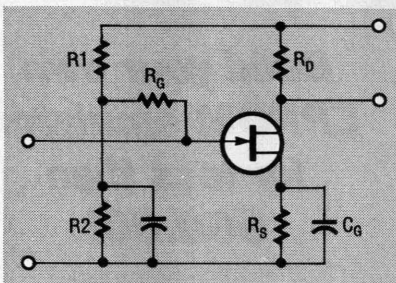


FIG. 11—JFET AMPLIFIER biased for low noise.

tion of frequency) vs. the device's noise current and voltage at the operating frequency, input impedance, and operating point. If the source impedance is less than a few hundred ohms, transformer coupling should be considered. Remember that if two FET's are in parallel, the effective input impedance will be halved. This doubles transconductance and noise current noise while halving the noise voltage.

Surprisingly, feedback has little effect on noise performance. Except for the possible introduction of noise from the feedback resistors, the addition of feedback does not affect the S/N ratio. Also, the amplifier configuration—common base, com-

mon emitter, or common collector—does not significantly affect device performance. This gives you the freedom to choose the amplifier configuration that best matches the signal source impedance.

The next step is biasing. Figure 9 shows a typical common-emitter amplifier stage. For a high-gain stage, R_C is unlikely to contribute noise, and noise produced by R_E is shunted to ground by C_E . Resistors R1 and R2 bias the base. For minimizing amplifier noise, the values for those resistors should be as small as possible.

Because these resistors have significant voltage across them, resistor excess noise must be considered. A tradeoff might be necessary. If calculations indicate that the excess noise is a problem, the circuit shown in Fig. 10 can be used. Noise from R1 and R2 is shunted by C_B . Only R3 contributes noise, but because the voltage drop across this resistor is low, excess noise does not present a problem. The analysis for a common-collector amplifier stage is similar to the common-emitter except that in this case the emitter resistor should be as large as possible to maximize the gain.

JFET biasing is illustrated in Fig. 11. Because of the low FET leakage current, R_G can be quite large so that it does not contribute much noise. Because a JFET's channel effectively isolates it from the gate, R_S and R_D are decoupled from the input. Therefore, noise from those resistors appears in the second stage, after the input signal has been amplified by the FET.

Once the input stage is designed, be sure that external noise is excluded. This means ensuring that the input signal source and input stage are well shielded and noise from the power supply is filtered out. Low-inductance capacitors can bypass power supply noise effectively. For more demanding applications, it might be necessary to use a separate modular power supply. In extreme cases, the isolation of a battery might be required.

R-E

AUDIO SCRAMBLING SYSTEM



Prevent eavesdropping by scrambling and descrambling your telephone conversations.

WILLIAM SHEETS AND RUDOLF F. GRAF

THERE ARE MANY INSTANCES where some form of speech encryption is needed for privacy or security. Complex and costly voice-scrambling systems are common in covert military operations, but simpler and less-expensive systems are adequate for discouraging the casual eavesdropper. The voice-encryption system described here inverts the frequency spectrum of the speech about a reference frequency to scramble the audio, and reinverts it to descramble the speech. Although the system is intended primarily to scramble telephone conversations, it is not limited to that. The device can also scramble tape recordings, which will be made intelligible only with the correct descrambler.

This method of speech scrambling is accomplished by mixing the audio input to be scrambled with a carrier tone as

shown in Fig. 1. The mixing process is carried out with a balanced modulator, which results in a double-sideband suppressed-carrier signal. The two resulting sidebands are the lower-sideband audio frequencies in the voice range (about 150–3000 Hz) and upper-sideband frequencies (about 3000–7000 Hz).

Since most voice circuits are designed for frequencies in the lower sideband range, the upper sideband is filtered out. The lower sideband contains frequencies that are similar to the original voice frequencies, but it has an inverted spectrum. Assuming a 3000-Hz carrier signal, an input signal of 500 Hz will produce a 2500-Hz output, and a 1-kHz signal will produce a 2-kHz output. The spectral energy of a human voice is more concentrated at the ends of the voice spectrum, mainly the 300–1000 Hz range, and some-

what less in the 2000–2500 Hz range. The resulting output will therefore have a very high-pitched sound, and be unintelligible. It can, however, be carried over normal telephone lines without being understood by eavesdroppers.

A digital voice-scrambling method is used in the circuit because it requires fewer parts than an analog system, needs no adjustment, and requires no switching. Because the descrambling process is the inverse of the scrambling process, the same circuit can be used for both functions. The encryption system has two channels for full-duplex operation, which allows easy two-way communication. Note that two complete systems—one at each end of a phone line—are required for two people to carry on a scrambled conversation.

The system operates as follows: An audio input is first fil-

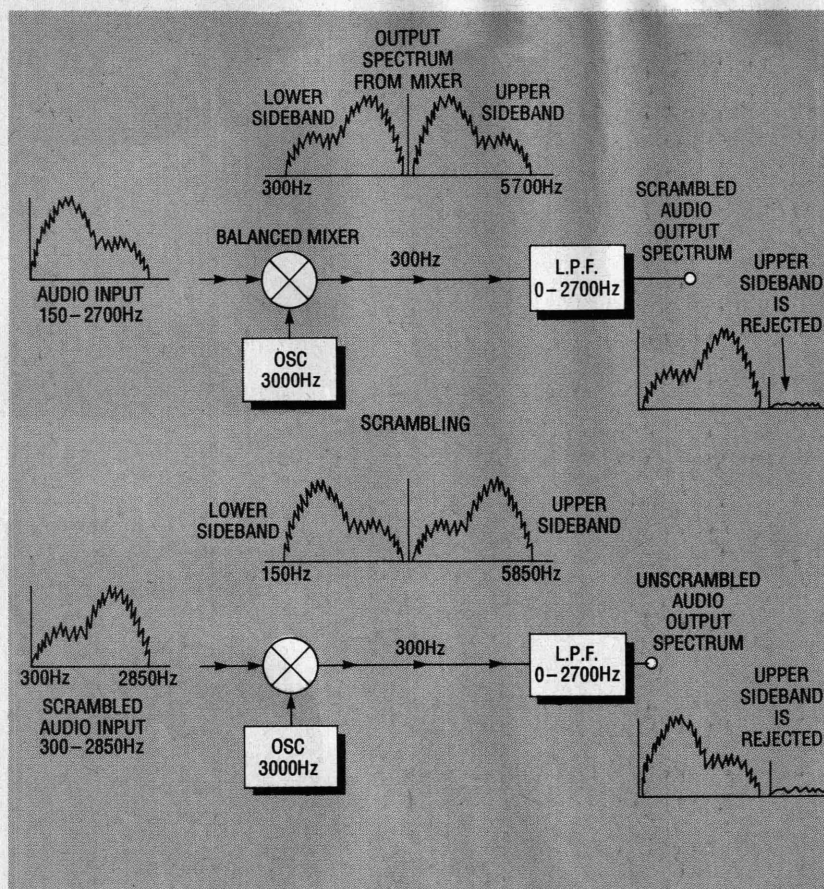


FIG. 1—AUDIO IS SCRAMBLED by mixing the audio input with a carrier tone. The two resulting sidebands are the lower sideband audio frequencies in the voice range (about 150–3000 Hz) and upper sideband frequencies (about 3000–7000 Hz).

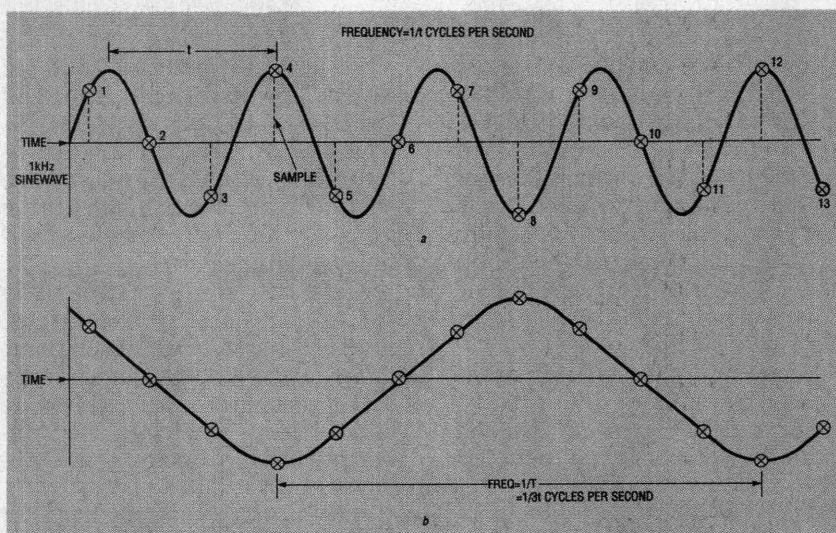


FIG. 2—A 1-KHZ SINEWAVE IS SAMPLED (a), and even-numbered samples are inverted, resulting in a lower-frequency sine wave (b).

tered with an active switched-capacitor bandpass filter to limit the frequency range to between 150 Hz and 2700 Hz. The signal is then digitized with a sampling rate of 5.86 kHz, which is more than double the

highest audio frequency (2700 Hz). Every second eight-bit digitized audio sample has its sign bit inverted while being fed to a digital-to-analog converter. That has the effect of inverting the spectrum of the analog out-

put signal after conversion from the digitized audio. The signal is then fed to a bandpass filter to remove switching components, leaving the final audio signal as one that corresponds to the input signal, except that its spectrum is folded around one-fourth of the sampling frequency, or 1465 Hz.

In Fig. 2-a it can be seen that

PARTS LIST

All resistors are 1/4-watt, 5%, unless otherwise noted.

R1—2.2 megohms
R2, R3—470 ohms
R4, R5—100 ohms
R6, R7—22,000 ohms
R8, R10, R14, R15, R17—1000 ohms
R9, R11—6800 ohms
R12—8200 ohms
R13—1000 ohms, potentiometer
R16—10,000 ohms, potentiometer
R18—R20—10,000 ohms
R21—33,000 ohms

Capacitors

C1—22 pF, NPO
C2, C3—82 pF, NPO
C4, C5, C11, C14, C16—0.01 μ F, disc
C6, C9, C10—1 μ F, 35 volts, electrolytic
C7, C8—10 μ F, 16 volts, electrolytic
C12—470 pF, disc
C13, C15—470 μ F, 16 volts, electrolytic

Semiconductors

IC1—74HC86 quad 2-input exclusive-or gate
IC2, IC7—74HC74 dual D-type flip-flop
IC3, IC4—74HC161 synchronous 4-bit binary counter
IC5, IC6—TP3054N Codec (National Semiconductor)
IC8—LM7905 -5-volt regulator
IC9—LM7805 5-volt regulator
D1, D2—1N4002 diode
Q1—2N3565 or 2N3904 NPN transistor

Other components

S1, S2—DPDT slide switch
J1, J2—RJ-11 4-conductor modular telephone jack
J3—1/8-inch power connector jack
XTAL1—3 MHz crystal (2.5 to 4 MHz usable)

Miscellaneous: PC board, 1/8-inch rubber grommets, insulated standoffs, case, hardware, 6–14 VAC, 100 mA transformer, IC sockets (optional), wire, solder.

Note: A kit of parts for one voice scrambler (two complete units are necessary) including a PC board and all parts that mount on it (does not include a telephone, case, phone cords, or wall transformer) is available from North Country Radio, P.O. Box 53, Wykagyl Station, New Rochelle, NY 10804 for \$59.00 + \$3.50 shipping and handling. A wall transformer is available for \$9.50. A North Country Radio catalog is \$1. New York State residents must add appropriate sales tax.

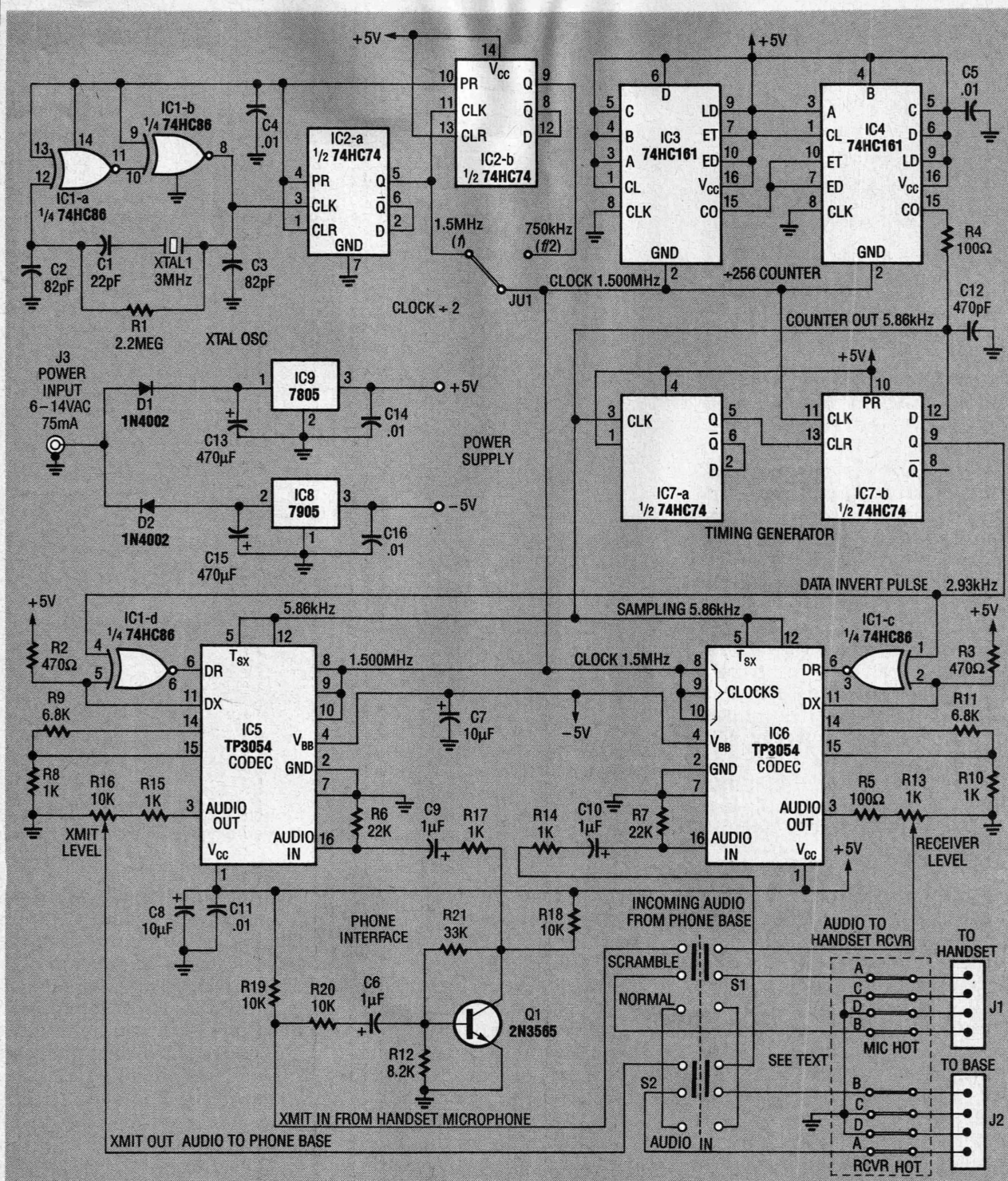


FIG. 3—VOICE SCRAMBLER/DESCRAMBLER SCHEMATIC. The clock and control circuitry supports the two CODECs, IC5 and IC6 (one for each channel).

a 1-kHz sinewave sampled as shown, with even-numbered samples inverted, results in a lower-frequency sinewave (b). The process also works in reverse; if the lower-frequency waveform (2-b) is sampled at the same points, and alternate samples inverted, the original waveform can be regenerated.

Circuitry

Figure 3 is the schematic of the voice scrambler/descrambler. Two chips—each a National Semiconductor TP3054 coder/decoder, or codec—form the heart of the circuit. The two integrated circuits, IC5 and IC6, contain all of the necessary A/D and D/A con-

verters, switched-capacitor filters, and associated tuning and control circuitry.

The scrambler's "ground" must be isolated with respect to true earth ground. Therefore the PC board of the scrambler should be mounted on insulated standoffs and fed about 75 milliamperes of isolated, low voltage AC from a wall-mounted

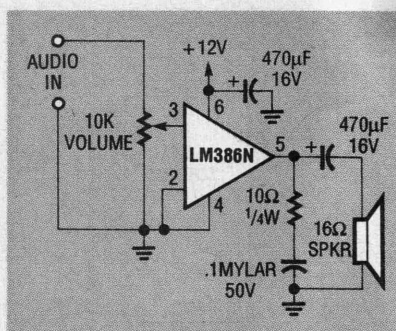


FIG. 4—FOR LOUDSPEAKER APPLICATIONS, a small audio amplifier like this one will generally be needed.

transformer. Do not connect the unit to the AC line without such a suitable isolating transformer.

Most of the rest of the circuit is clock and control circuitry that supports the two codecs. The clock signal is generated by an oscillator made up of 3-MHz crystal XTAL1 and IC1-a and -b. The 3-MHz signal is divided in half by IC2-a to produce the main 1.5-MHz clock signal, and IC2-b again divides by 2 to produce an optional clock frequency of 750 kHz. That signal is further divided down by IC3 and IC4 to produce 5.86- and 2.93-kHz signals. D-type flip-flop IC4 produces a 2.93-kHz pulse train that's used for bit-stream inversion.

The 5.86-kHz pulse shifts a serial data stream, eight clock pulses wide, from the codec's A/D converter to the D/A converter. Data from an A/D converter (pin 11 of IC5 or IC6) is fed to IC1-d or -c, respectively. Those EXCLUSIVE OR gates act as inverters if one input is held high, or as straight-through non-inverting buffers if the other input is held low. By applying a 2.93 kHz pulse on one input, alternate data-stream sign bits (which occur at a 5.86-kHz rate) are inverted. Therefore, the data from pin 11 of IC5 (or IC6) that is fed back to the D/A converter section (pin 6) has every other sample reversed in sign. That has the aforementioned effect of inverting the frequency spectrum of the reconstructed analog signal.

The circuitry required to interface the voice-encryption system to a telephone is contained

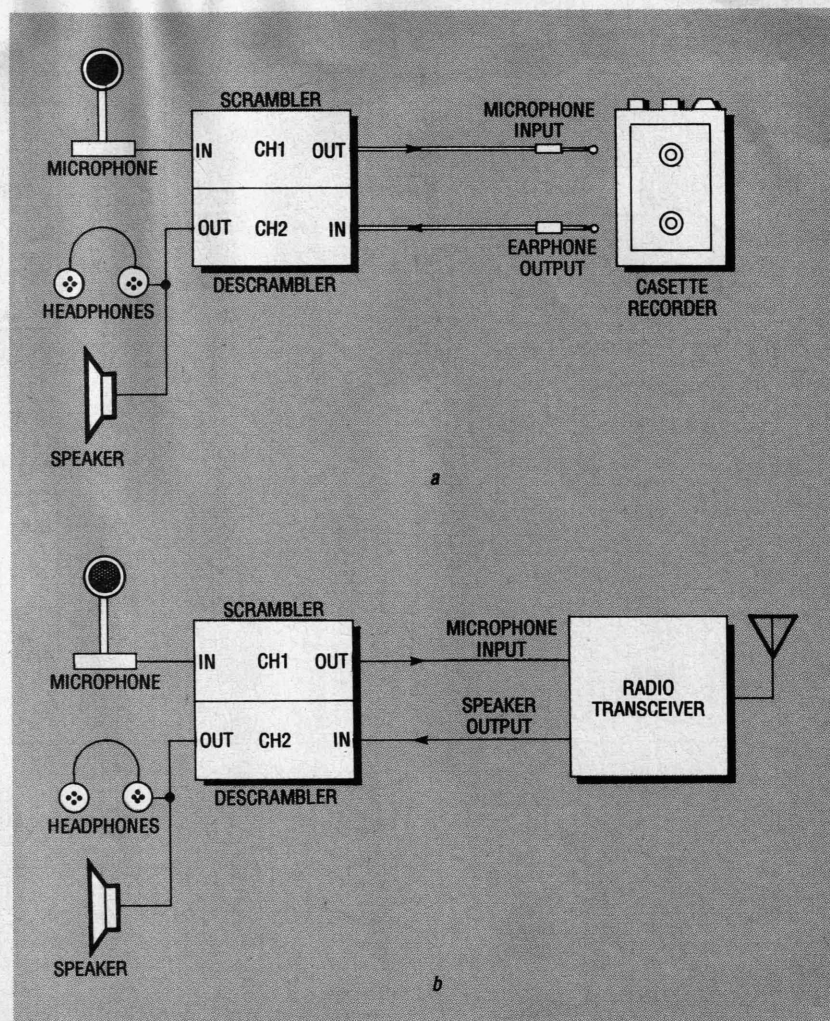


FIG. 5—THE SYSTEM can make scrambled audio recordings (a), or it can also be fitted to a radio transceiver (b).

in Fig. 3. Resistor R17 couples audio from amplifier Q1 to IC5. Transistor Q1 provides about a 10 dB voltage gain.

Modular jacks J1 and J2 connect to S1 and S2 via jumpers that are configured for your particular telephone set. Because direct insertion of the device in a telephone line would not be feasible without a lot of switching due to ringing and signaling considerations, it is necessary to install this device in the handset line. This way only the microphone and earphone have to be considered.

The TP3054 can drive a 600-ohm load (the impedance of a telephone line) directly. If telephones are not being used, simply use the input and output pins directly of each codec. To have the chip drive a loudspeaker, a small audio amplifier, such as that shown in Fig. 4 is

required. Note that, when using a microphone to input audio to the codec, some microphones have internal audio amplifiers and can produce well over one volt of audio. Those microphone outputs can be input directly to the codec. Low-output microphones require amplification.

A switching network (S1 and S2) is added on the PC board to switch the scrambler in or out of the telephone circuit. Resistor R13 sets the sound level at the telephone receiver, and R16 is set for optimum reception at the other end of the telephone line.

Figure 5 shows two more applications; 5-a shows how the system can be used to make scrambled audio recordings, and 5-b shows how a radio transceiver can be fitted with this device. (Bear in mind that in services such as amateur ra-

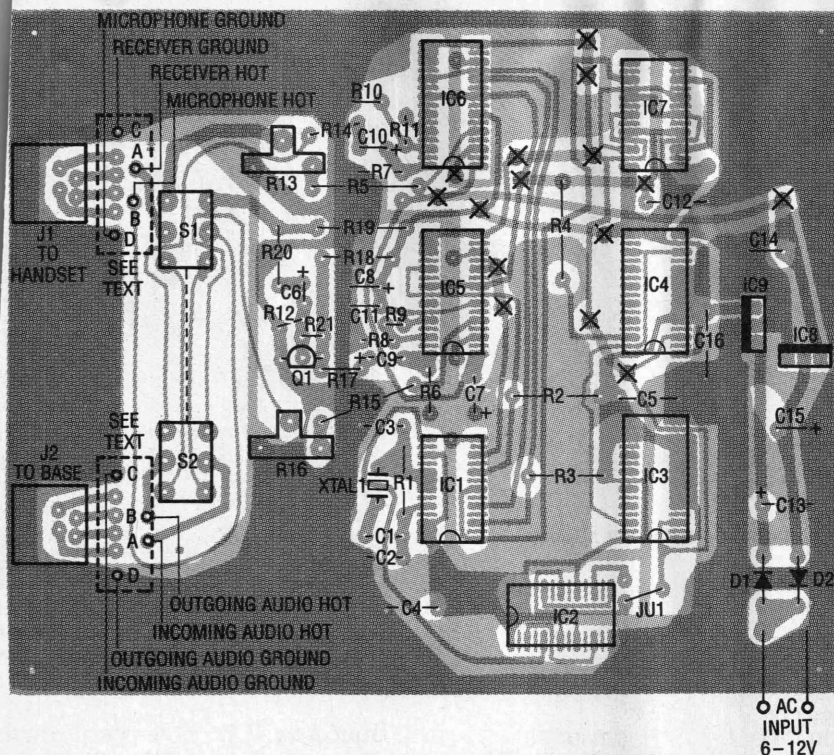
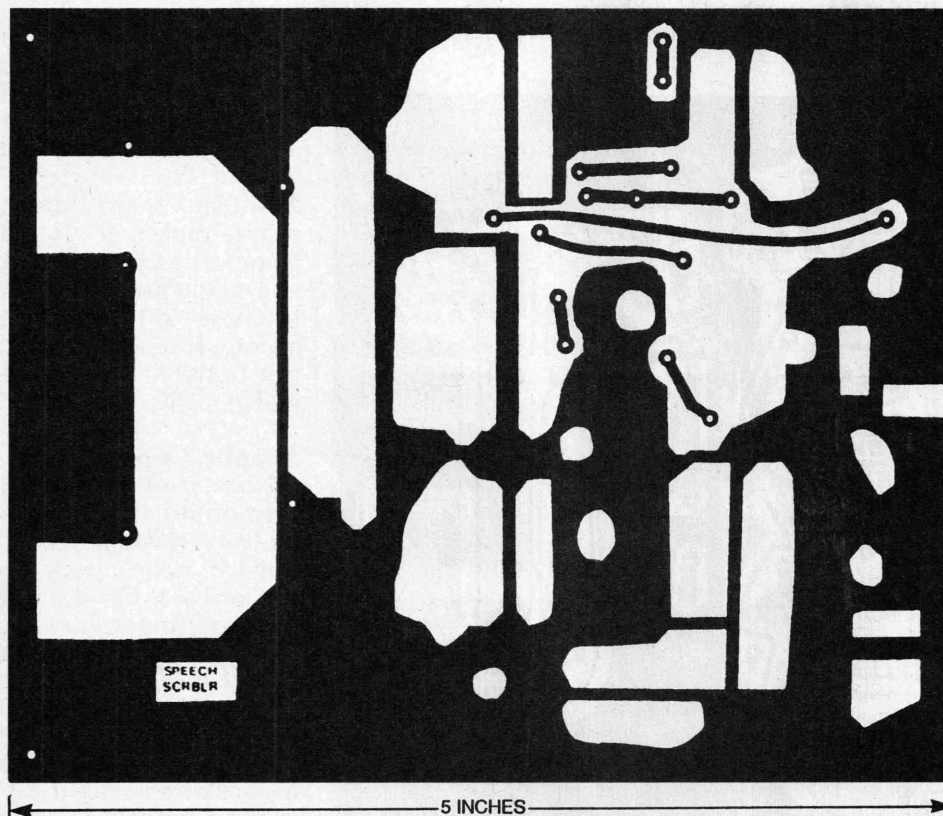


FIG. 6—PARTS-PLACEMENT DIAGRAM. Install through-board jumpers in all locations marked with an "X" and solder them on both sides of the board. All component leads that pass through a hole copper clad on both sides must also be soldered on both sides.



COMPONENT-SIDE FOIL PATTERN for the voice scrambler.

Construction

The circuit can be point-to-point wired by hand or made on the double-sided PC board for which foil patterns are provided in this article. Figure 6 is the parts-placement diagram. Because the PC board does not have plated-through holes, first install through-board jumpers in all locations marked with an "X," and solder them on both sides of the board. All component leads that pass through a hole with copper on both sides must also be soldered on both sides.

Install all passive components such as resistors, capacitors, jumpers, and switches on the PC board. Then install voltage regulators IC8 and IC9. Also install the sockets for the rest of the ICs, if you are using them (they are recommended). Before the ICs are inserted in their sockets, apply 6 to 12 volts AC to the junction of D1 and D2 and ground. Check for 5 volts at pin 4 of IC5 and IC6, and pin 16 of IC2, IC3, and IC4, and pin 14 of

IC1. Next, verify -5 volts at pin 1 of IC5 and IC6. If these voltages check out, insert the ICs in their sockets. Figure 7 shows how to gang switches S1 and S2 together mechanically, and Fig. 8 shows the finished board, with the ganged switches, mounted in a case.

Testing

Verify that a 5-volt peak-to-peak, 1.5-MHz signal exists at pin 2 of IC3. Check for a 5.86-kHz pulse train at pin 15 of IC4, pins 5 and 12 of IC5, and IC6. Check for 2.93-kHz pulse train at pin 1 of IC1-c and pin 4 of IC1-d. Due to the short pulse width (250 nanoseconds), it might be difficult to see these pulses with an economy model oscilloscope.

If all checks out, apply a 0.5-volt peak-to-peak, 1-kHz tone to the junction of R17 and C9; a 2-kHz tone should be pro-

dio, it is illegal to use encryption, so check FCC rules to

verify the legality for any intended application.)

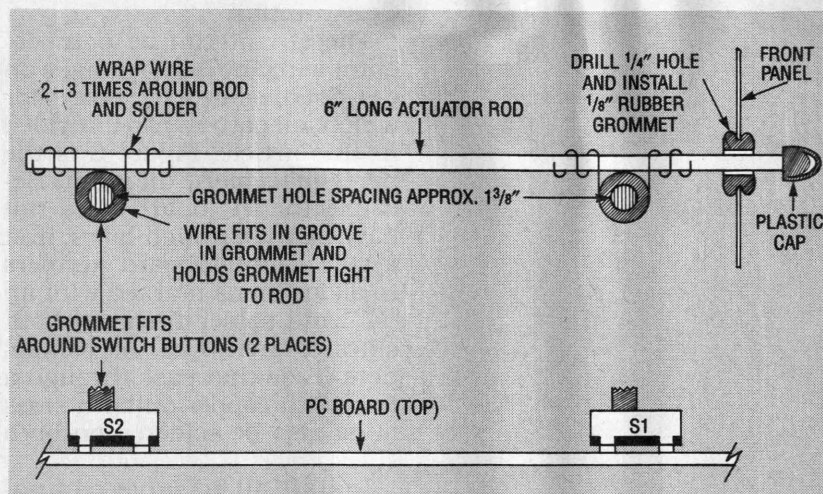


FIG. 7—SWITCHES S1 AND S2 must work in conjunction with one another, so they must be ganged together as shown here.

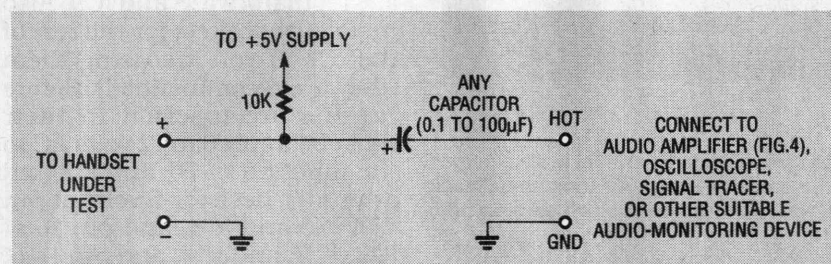
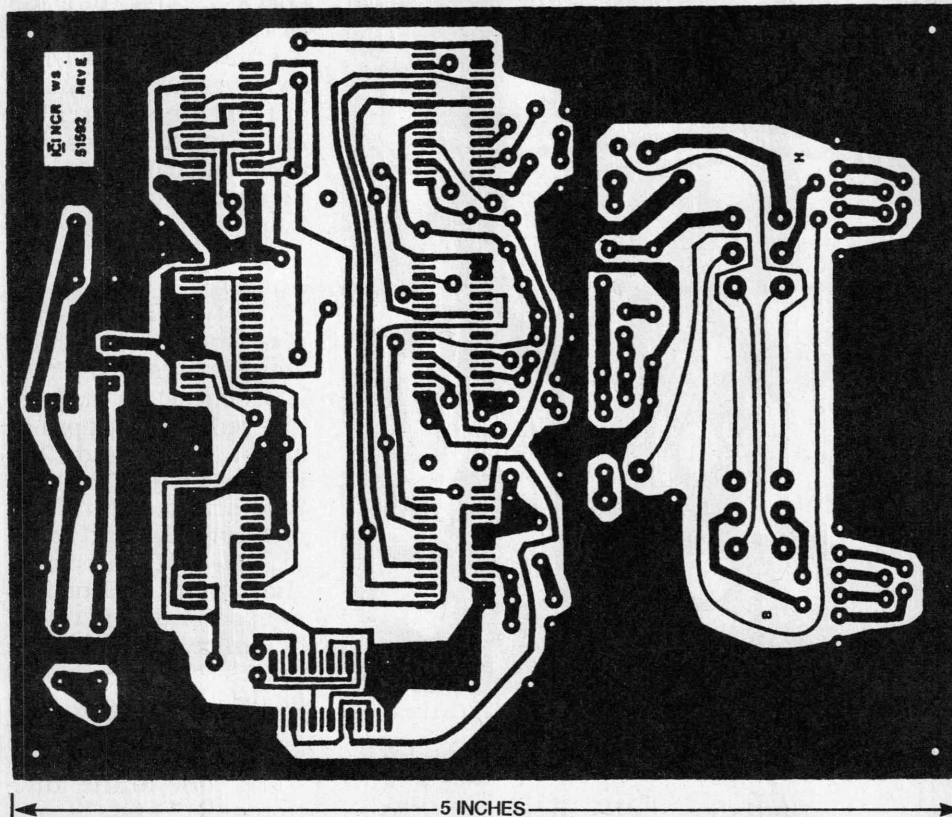


FIG. 9—TO DETERMINE THE POLARITY of the microphone leads, connect the microphone pair to this test circuit and see if the microphone works; if not, reverse the connections.



SOLDER-SIDE FOIL PATTERN for the voice scrambler.

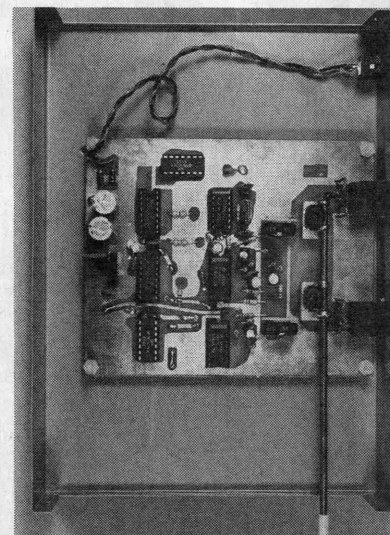


FIG. 8—HERE'S WHAT THE FINISHED board looks like mounted in the case. Notice how S1 and S2 are ganged together.

duced by IC5. Now temporarily connect pin 3 of IC5 to the junction of C10 and R14 using a 100K resistor. Pin 3 of IC6 should produce the original 1-kHz tone.

Next, apply an audio signal (from a tape deck or radio) with about 2 volts peak-to-peak to the junction of R17 and C9. Listen to the output at pin 3 of IC5; it should sound "scrambled." Now listen to the output of channel 2. It should be normal, but note that the high and low frequencies might sound somewhat attenuated due to the narrow bandwidth of the system.

Adapting a phone

Note that this unit cannot be connected directly to the phone lines. It will handle speech audio only, and will not pass ringing signals or rotary dialing pulses. It will also distort dialing tones. You must use only a phone whose handset has accessible microphone and receiver connections. You cannot use a unified telephone (where the dial or pad is built into

Continued on page 46

you decide to make boards on a regular basis. You can buy an etching tank from suppliers that carry materials for making PC boards, or even build one (see **Radio-Electronics**, December 1989).

When you can see that the unwanted copper is almost fully etched, take the board out of the etchant and plunge it into the bucket of water to rinse off the etchant. If it is not etched enough, put it back into the etchant. It's better to stop etching too soon and then put it back in again than to etch it too much. When the board is done, rinse it off and towel dry it.

If your board does not etch completely, even with fresh developer and etchant, you must repeat the entire procedure, exposing the board a bit longer to the sun lamp. However, if traces are being etched away along with the unwanted copper, then you must reduce the exposure time. Once you determine the correct exposure time, it will work every time. Call your local

city officials to find out how to properly dispose of the used chemicals—don't just pour them down the drain!

With a piece of fine steel wool, polish the board to remove all remaining etch resist. Inspect the board for any broken traces. You are now ready to drill the board. Remember to wear safety goggles. Although a drill press is best for drilling PC boards, a hand drill and a steady hand will also work.

When double-sided PC boards are professionally made, holes are plated through the board to connect traces from top to bottom. With our technique there is no plating through of the holes. We simulate the through-holes by inserting short lengths of wire through the holes and soldering them on both sides. In through-holes where component leads pass through, the leads are simply soldered on both sides. Because some components have leads that are difficult to solder on the component side, it is good practice to

place as many traces as possible on the solder side when laying out the board.

When your board is complete you can assemble your prototype. If everything checks out you are ready to send the information to your PC board manufacturer and have them make as many boards as you need with the assurance that they will work properly.

SOURCES

Everything you need to make your own PC boards can be obtained from the following suppliers:

- Kepro Circuit Systems, 630 Axminster Drive, Fenton, MO 63026-2992 (800) 325-3878, (314) 343-1630
- Datak Corporation, 3117 Paterson Plank Road, North Bergen, NJ 07047 (201) 863-7667.
- A list of circuit-board houses who provide prototyping and full production services for boards up to 16 layers can be obtained free of charge from Skychaser, 980 Sherwood Place, Eugene, OR 97401 (503) 345-4609.

AUDIO SCRAMBLING

continued from page 42

the handset). The handset should preferably have an electret microphone. However, carbon microphones (found in older phones) can be used, if necessary, but R19 should be changed to about 1K, and R20 might have to be increased if excessive audio from a carbon microphone overdrives Q1, causing distortion.

Depending on the phone you have, you must make the proper jumper connections on the PC board near J1 and J2. A Radio Shack telephone model No. ET-171 (cat No. 43-374) was used in the author's prototype.

You must identify the following things on your phone(s):

1. The handset microphone and earpiece connections
2. The type of microphone (electret, dynamic, or carbon)
3. Microphone polarity (if it's the electret type)
4. The base connections

There are usually four wires that connect a telephone handset to its base. If you can't visually identify the wires after disassembling the handset, try connecting a 1.5-volt battery to alternate pairs of wires on the handset until you hear a click in the earpiece. Mark these as the receiver leads; there should be between 50 and 1000 ohms between them. The other two leads are for the microphone.

Check for short circuits between both of the receiver leads and the microphone leads with an ohmmeter on a high resistance range. A low resistance or a short between any two leads indicates that they are the ground leads for the microphone and receiver. If you find no continuity between the two sets of leads, connect the microphone pair to the test circuit shown in Fig. 9, and see if the microphone works; if not, reverse the connections. This will identify the microphone's hot and ground leads. Once you've identified all of the handset

leads, note their positions on the modular connector. The telephone base connections can now be determined from the positions of the handset leads at the modular connector.

When you have all of the telephone connections identified, install the jumpers on the PC board near jacks J1 and J2. In Fig. 6 there are four jumper pads labeled A-D near each modular jack; the pads are also labeled by function. Once you know the signal positions at jacks J1 and J2 for your phone, install four jumpers per jack to properly route the signals.

The finished board can be mounted in a case like the one pictured in Fig. 8, or in any other suitable case. The case pictured allows the telephone to be placed on top of the scrambler without taking up any extra space.

With a pair of scrambler phones in hand, you're ready to start talking. All you need now is someone to talk to and a confidential topic to discuss. Ω

BUILD A MICRO TV TRANSMITTER

This wireless mini video transmitter sends a picture to your TV set. It's ideal for security monitoring or observing nature.

MARTIN MCKINNEY & GARNET BRACE



THE MINIATURE SINGLE-BOARD VIDEO camera used in this project is an excellent example of how recent advances in integrated circuit and surface-mount technology have led to remarkable advances in video technologies. A complete, reasonably priced solid-state camera will easily fit in the palm of your hand.

This article describes how you can build a simple modulator/transmitter and connect it to a commercially available video camera to obtain a portable, battery-operated camera that can transmit black and white pictures from remote locations back to your TV set. No antenna is needed to transmit to a nearby TV set, but if an external antenna is used, the modulator has enough power to transmit about 100 yards. The RF modulator/transmitter can transmit over standard NTSC channels 7 to 13 or UHF channels 14 to 29.

The mini video camera on which this project is based measures $0.91 \times 1.81 \times 2.76$ inches

and it weighs only 1.3 ounces. The camera requires a 6- to 12-volt DC power supply, and draws a current of only 65 milliamperes.

The applications for a battery powered, hand held TV transmitter are limited only by your imagination. Home or business security come most readily to mind. Because cables are unnecessary, there is no danger of an intruder cutting cables and disabling the camera. The camera can be placed in an unobtrusive location protected from the weather.

You can also use this camera transmitter to keep an eye on an infant in a crib or observe a disabled or bedridden person. And you could also put this camera to work as the "eyes" of a robot or for monitoring industrial welding or machining operations that would pose a safety threat to people standing too close.

If you are a farmer, you can monitor the behavior of live-

stock in a barn, and if you're a nature hobbyist, you can observe wild animals and birds at close range. The ability of the camera to "observe" action in reduced light will be particularly attractive.

The camera/transmitter system can be powered from an AC to DC wall-mounted adapter, or it can be powered from a standard 9-volt battery. The service life of an alkaline-manganese battery is approximately four hours.

Camera modules with several different configurations are available from the source given in the parts list. One is a super wide-angle (110° field-of-view) 3-mm, $f1.8$ lens; the other is a narrower angle (78° field-of-view) 4.3-mm, $f1.8$ lens. A 12-mm, $f1.8$ lens is available for those interested in aerial or nature photography. This camera, when placed behind the viewfinder of a single-lens reflex (SLR) still camera, will accept the full field-of-view of the still camera lens and permit remote

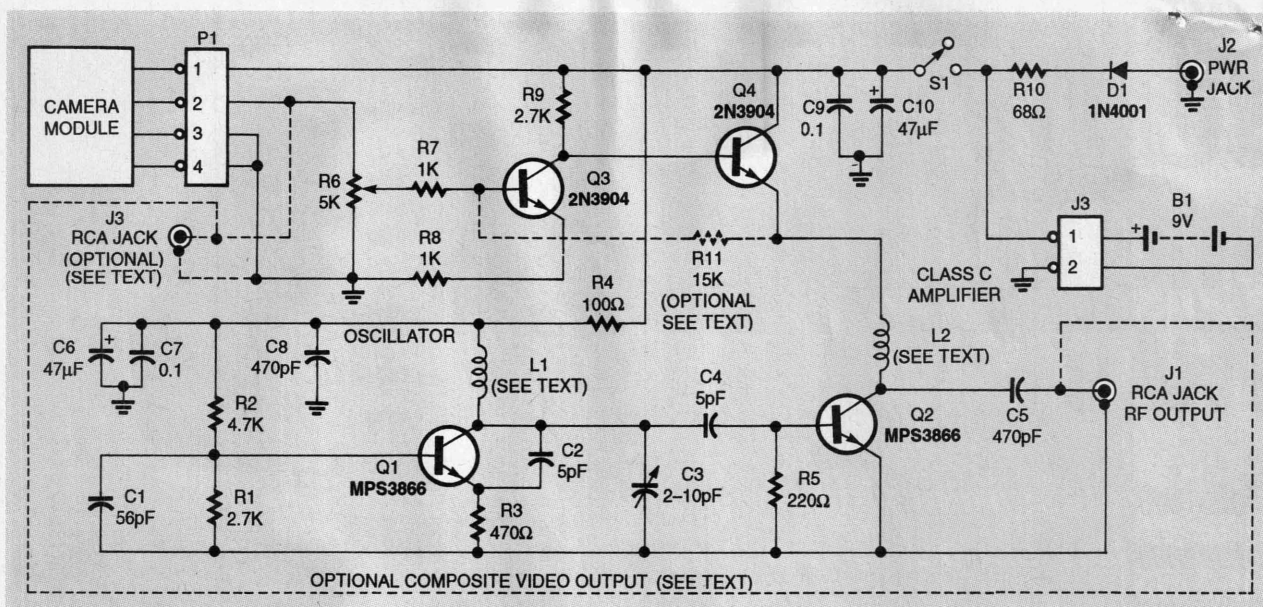


FIG. 1—MODULATOR SCHEMATIC. Coil L1 is the key component that determines the frequency of oscillation.

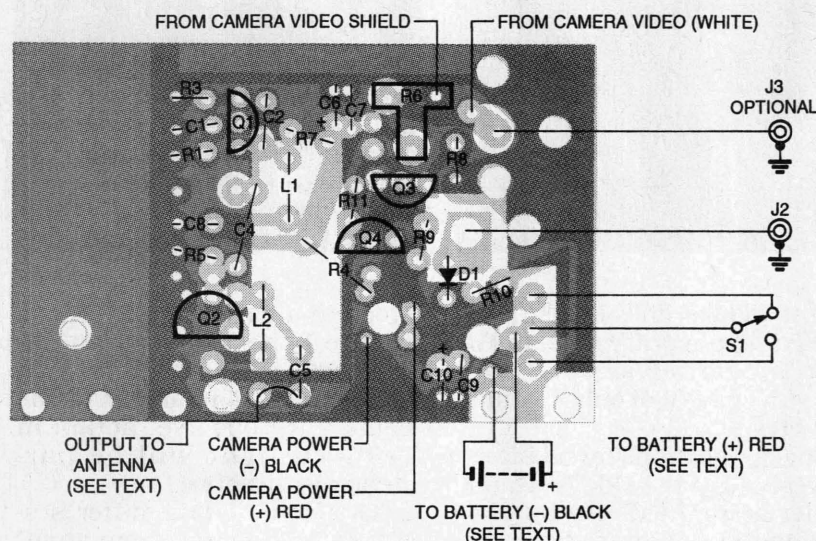
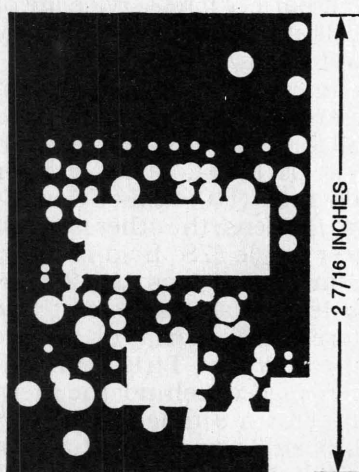
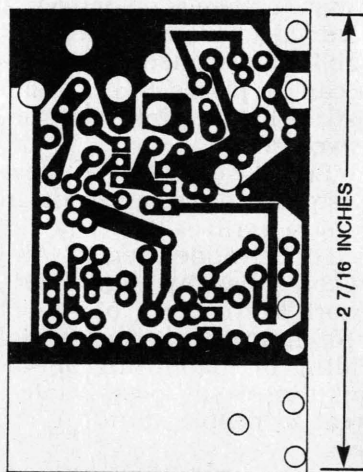


FIG. 2—PARTS-PLACEMENT DIAGRAM. Note that the board can be cut, if needed, to fit in a smaller case.



SOLDER SIDE of the board.



COMPONENT SIDE of the board.

monitoring of the picture to be taken.

The CCD mini camera is also sensitive to infrared (IR) radiation, and when paired with a simple infrared light source, can be used to see in the dark. For robotic or industrial applications, a single infrared-emitting LED emits enough energy to illuminate nearby objects.

The modulator is built on a $2\frac{1}{4} \times 1\frac{1}{2}$ -inch printed-circuit board, which is mounted on insulating standoffs above the camera module. A wiring harness and plug connect the camera to the modulator.

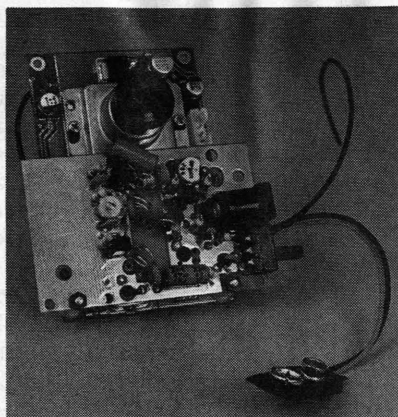
Modulator operation

The schematic of the transmitter is shown in Fig. 1. The circuit operates as follows: Transistor Q1 is the amplifier in a transistor oscillator that oscillates in the frequency range of VHF channels 7 to 13 (174 to 216 MHz) or slightly above (250 MHz), depending on the desired transmitter frequency. The output from the resonant tank circuit of the oscillator, formed by coil L1 and capacitor C3, is capacitively coupled by C4 to the base of the Class-C amplifier stage that includes Q2.

The combination of L2 and the parasitic capacitance of the PC board broadly tunes the output of Q2 to either the same range of frequencies as the os-

cillator stage, or to a frequency that is twice that of the oscillator stage, which permits the transmitter to cover Channels 14 to 29 (470 to 580 MHz).

The Class C amplifier stage is collector-modulated by a two-transistor modulator that consists of Q3 and Q4. The output from the camera is directly coupled to Q3 through variable resistor R6. The ratio of resistors R9 and R8 in the collector and emitter circuits of Q3 determines the gain of approximately three for this stage. The output of the camera is an NTSC composite video signal of approximately 2 volts peak-to-peak amplitude with a 0.7 volt DC offset. The output of Q3 is directly coupled to emitter follower Q4, which collector-modulates the Class C output stage.



THE MODULATOR BOARD after being mounted on the video camera module.

The modulator has been designed to work primarily with the camera described, but it can also be used as a modulator for any composite video device. This is accomplished by chang-

ing the value of R9 from 2700 ohms to 4700 ohms, thus increasing the gain of the input stage and adding 15 K resistor R11, which removes the DC offset from the amplifier stage. Provision has been made on the modulator printed circuit board to add an optional right-angled RCA connector that can provide baseband video output or accept baseband video input when the modulator is used with other devices.

Diode D1 and resistor R10 form a simple battery-charger circuit. Diode D1 also protects the camera from reverse voltage should a charger with the wrong polarity be connected accidentally.

Construction

A complete modulator circuit board can be purchased from the source given in the parts list. However, foil patterns are provided for those who want to make their own. Note that the circuit board stock is 0.031-inch thick glass-epoxy with copper foil on both sides. (Paper-based PC boards should not be used at the high frequencies produced in this circuit).

Certain parts of the ground-plane foil on the component side of the board must be removed to provide insulated surfaces for some components. See the parts-placement diagram in Fig. 2.

Mount all components on the non-solder ground-plane side of the board. Mount all resistors vertically after forming the leads. Mount all disc capacitors as close as possible to the ground plane. Both electrolytic capacitors are mounted horizontally and their axes lie parallel to the surface of the board.

Begin by mounting the shortest components first. Start first with the variable capacitors, variable resistor, and the ceramic disc capacitors. Then add the transistors, resistors, and coils.

Although there are no critical component positioning requirements, take special care when installing the transistors. They look alike, but the circuit will not work if they are acciden-

PARTS LIST

All resistors are 1/4-watt, 5%, unless otherwise specified.

- R1, R9—2700 ohms (see text)
- R2—4700 ohms
- R3—470 ohms
- R4—100 ohms
- R5—220 ohms
- R6—5000 ohms, potentiometer
- R7, R8—1000 ohms
- R10—68 ohms
- R11—15,000 ohms (see text)

Capacitors

- C1—56 pF, ceramic disk
- C2, C4—5 pF, ceramic disk
- C3—2 to 10 pF trimmer
- C5, C8—470 pF, ceramic disk
- C6, C10—47 μ F, 16 volts, aluminum electrolytic
- C7, C9—0.1 μ F, ceramic disk

Semiconductors

- Q1, Q2—MPS3866 NPN transistor
- Q3, Q4—2N3904 NPN transistor
- D1—1N4001 rectifier diode

Other components

- L1—3 turns tinned bus wire wound on a 3/16-inch drill bit for channels 7–13 OR 2 turns tinned bus wire wound on a 3/16-inch drill bit for channels 14–29 (see text)
- L2—Same as L1 (see above)
- J1—Bulkhead-mounted RCA jack (optional)
- J2—2.1 mm DC power jack
- J3—right-angle RCA jack (optional)
- S1—SPDT miniature switch

Miscellaneous: CCD Camera Module (Chinon CX-102), PC Board, plastic case, 9-volt al-

kaline battery, hardware, 5/16-inch nylon standoffs, 1/2-inch No. 2-56 screws and nuts, coaxial cable, wire, solder.

Note: The following items are available from Bayview Hobby Supplies, 14800 Yonge St. #110, Aurora, Ontario, Canada L4G 1N3. Tel: (416) 713-0528. Fax: (416) 841-4436:

- Camera module including wiring harness—\$229.00

- (specify Model M100W with wide angle 3 mm lens and 110 degree field of view or Model M100N normal angle 4.3 mm lens and 78 degree field of view)

- Modulator PC board (M200)—\$12.95

- Drilled plastic case (M201)—\$14.95

- 9-volt AC wall adapter (M202)—\$14.95

- Modulator kit (M300)—\$49.95

- Completed modulator (M301)—\$99.95

- Kit including camera module, all parts, hardware, and case (M400)—\$299.00

- Semi-kit, includes camera module and fully assembled and tested modulator board (M401)—\$329.00

- Completely assembled and tested camera and modulator mounted in case (M402)—\$349.00

All prices in U.S. funds.

tally interchanged. Capacitor C8 is an RF bypass capacitor that forms part of the tank circuit. Its leads should be as short as possible.

Coil L1 is a 2- or 3-turn coil (depending on the output frequency selected) of No. 22 AWG tinned wire wound on an $\frac{3}{16}$ -inch drill bit. The coil, in conjunction with variable-capacitor C4, determines the oscillator frequency. This frequency is in the range of 174 to 216 MHz for channels 7 to 13, or 235 to 280 MHz for channels 14 to 29. (The oscillator frequency is doubled to 470 to 560 MHz in the Class C stage for the higher channels). Diode D1 should be mounted with its anode closest to power jack J2.

After all components are soldered, install the battery clip, wiring, and plug for the TV camera module. These wires enter from a hole in the bottom of the board to provide strain relief, and they are soldered after insertion into holes in the top of the board. The modulator can be operated without an antenna but, if longer range is desired, an antenna can be connected to output capacitor C5 (see options section).

Alignment

When the board components have been assembled and soldered, check the current drain with the modulator connected to a 9-volt supply. The current drain should be about 25 milliamperes, and should drop to about 10 milliamperes when the main oscillator is inhibited (which is accomplished by touching L1 with a finger.) This simple test indicates that the oscillator is functioning and the Class C stage is amplifying.

After testing, mount the finished modulator board above the camera module with two $\frac{5}{16}$ -inch long nylon standoffs and two $\frac{1}{2}$ -inch long No. 2-56 screws and nuts.

After mounting the modulator board, the camera plug can be connected. At this time, check the current drain again. With the camera connected, the drain from a 9-volt supply should be about 85 milliamps.

With a television receiver tuned to the desired channel near the completed unit and power applied to the camera and modulator, tune C3 until a video image appears on the screen. Turn R6 clockwise to a position that gives a picture with maximum contrast and no "tearing."

If no video image appears on the screen after all other possible sources of assembly error have been eliminated, adjust L1 by separating its turns to raise the output frequency or compress the turns to lower the frequency. When using video sources other than the camera, it will be necessary to change R9 from 2.7K to 4.7K to increase the gain of the modulator. Also add the 15K resistor R11 to eliminate DC offset.

Once these adjustments are made, mount the completed unit in the plastic case and fasten the back in position with four self-tapping screws.

There is no voltage regulation to compensate for battery-voltage variation. The resistor values in the oscillator section have been selected so that the oscillator exhibits a minimum frequency shift over the usable battery range.

Options

Provision has been made in this design to offer the user a number of options to give the camera extra versatility.

• **Frequency.** The modulator is designed to operate either on Channels 7 to 13 or 14 to 29. For Channels 7 to 13, coils L1 and L2 are three turns of No. 22 wire wound on a $\frac{3}{16}$ -inch form. For Channels 14 to 29, coils L1 and L2 are two turns of No. 22 wire wound on a $\frac{3}{16}$ -inch form.

• **Enclosure.** You might want to mount the unit in its own enclosure such as a picture frame or lamp for surveillance. The modulator board can be mounted to the camera without modification, or it can be cut with shears along line A-A to make it smaller.

• **RCA Input/Output Jack.** Provision has been made on the PC board for a right-angle RCA jack. (Note that the board will

not fit in the case pictured if this option is selected.) This option provides composite baseband output from the RCA jack or composite video can be introduced to the jack from an external source.

• **Wall Adapter.** The camera unit can be powered from any negative ground (positive tip) 9-volt wall adapter. When a wall adapter is used, the battery should be removed. Be sure that the output of the adapter does not exceed 12 volts.

• **Rechargeable Battery.** A rechargeable nickel-cadmium battery can replace the suggested alkaline battery, although battery life will only be about half as long (two hours instead of four hours) before the battery needs a recharge. Resistor R10 has been selected to recharge the battery in approximately four hours. Many rechargeable 9-volt batteries actually have terminal voltages of 7.2 volts and will charge at a higher rate. Most wall-mounted chargers are poorly regulated and their voltage rises dramatically when lightly loaded. Make sure that the battery does not overcharge.

• **RF and Baseband Output.** For short range use (up to 30 feet), no antenna is necessary as long as there is an adequate antenna on the TV receiver. For longer range, a 1 foot length of No. 18 stranded copper hookup wire can be connected to output capacitor C5 and extended through the case. If you want both baseband output and RF output, an optional RCA jack can be mounted on the front panel of the case and the output from C5 can be connected to it by a short length of wire.

A length of coaxial cable from the camera harness can be connected from the jack to the camera output on the modulator board. (The cable effectively blocks RF from getting back into the camera.)

If you want baseband output, an RCA plug and cable inserted into the jack will provide composite video output to a video monitor. If you want RF output, an antenna can be inserted into the RCA jack. Ω

THE SPECTRUM



ANALYZER

Vertical pulse height represents signal strength and horizontal pulse position represents frequency. Learn more about this important but frequently overlooked instrument

FERNANDO GARCIA VIESCA

THE SPECTRUM ANALYZER LETS YOU view the frequency domain the same way the oscilloscope lets you view the time domain. This article is a primer (or refresher) on the spectrum analyzer, an extremely powerful and versatile test instrument that is frequently overlooked by many who have never taken the time to explore its special properties.

Long considered too specialized and too expensive for general use, the spectrum analyzer has recently become an important test instrument in cable television, computer networks, digital radio, and automated component testing. However, for years it had been a well established instrument in radio and radar transmitter design and maintenance.

Government requirements mandating the control of electromagnetic interference (EMI) have made the spectrum ana-

lyzer a factory-production and quality-control tool in the manufacture of many different kinds of electrical and electronic equipment.

The selection of the appropriate spectrum analyzer for a given task will depend on such factors as the frequency sweep rate, frequency sweep range, bandwidth, and the center frequency of its intermediate frequency system. Other considerations are the bandwidth of its video amplifier and the analyzer's overall sensitivity.

Spectrum analyzers are useful in the analysis of continuous-wave (CW) signals, amplitude-modulated (AM) signals, frequency-modulated (FM) signals and pulse-modulated signals: pulse-amplitude modulated (PAM), pulse-width modulated (PWM), and pulse-position modulated (PPM) signals.

Before discussing the spec-

trum analyzer and how it works, a review of some of its basics will be useful. Figure 1 illustrates two different ways of looking at the same signal. The waveform viewed at the left is an instantaneous voltage plotted against time, as seen on an oscilloscope. The view at right shows the same waveform, but here the voltage is positioned with respect to frequency rather than time.

A complex signal has more than one frequency component (spectral line), which indicates the amount of energy that is present at each frequency. The more complex the waveform, the greater the number of spectral lines.

According to the French mathematician and physicist Jean Fourier, any time domain electrical signal consists of one or more sine waves, each with its own frequency, amplitude,

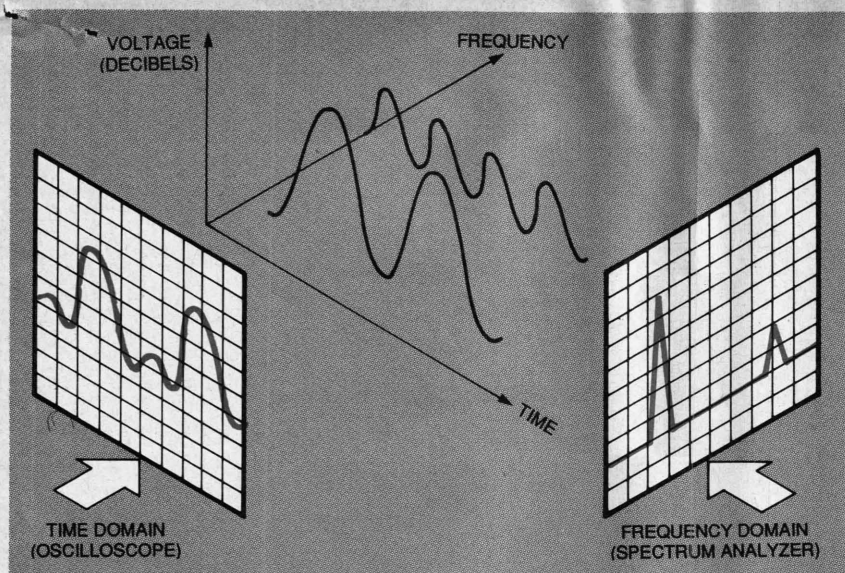


FIG. 1—TIME-DOMAIN MEASUREMENTS made with an oscilloscope (left) are compared with frequency domain measurements that can be made with a spectrum analyzer (right).

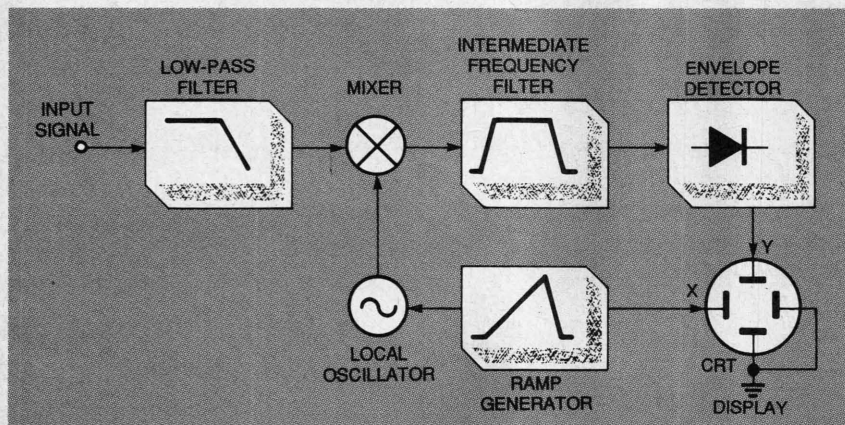


FIG. 2—SIMPLIFIED SPECTRUM ANALYZER BLOCK DIAGRAM. The input signal is mixed with local oscillator (LO) signal in the mixer. The intermediate frequency (LO) is filtered, detected and displayed on the CRT's vertical axis. The ramp generator triggers the CRT's time base and the LO.

and phase. As a result of his insight, it was found that any complex time-domain waveform can be broken down into separate sine waves that can be evaluated separately. In effect, any time domain signal can be transformed into its frequency-domain equivalent.

Theoretically, any signal must be evaluated for an infinite length of time when making a transformation from the time to the frequency domain. However, in practice, it is assumed that the behavior of a signal for time intervals ranging from milliseconds to a minute will give a reasonably accurate indication of the waveform's long-term characteristics.

Time vs. frequency domain

Time-domain measurements from an oscilloscope provide an overall view of the signal waveform, including such characteristics such as pulse rise and fall times, overshoot, and ringing. By contrast, the spectrum analyzer provides a panoramic display of the signal distribution in a selected portion of the radio frequency or microwave bands. The display can show the presence or absence of signals of interest, their frequencies or harmonic content, frequency differences, relative amplitudes, and the modulation form, if any.

In the United States and most of the advanced countries of the

world, radio equipment must meet government-imposed standards for spectral purity. The spectrum analyzer permits the viewing of the bandwidth of the modulated signal, and it also gives an immediate indication of the harmonic content of a transmitter's signal.

Harmonics of the carrier signal might interfere with other systems operating at the same frequencies as the harmonics. Improper RF transmission conditions such as splatter or over modulation can be easily seen as excessive bandwidth. These deviations can be observed on a spectrum analyzer. This is helpful in making adjustments and corrections.

How does it work?

There are four basic spectrum analyzer architectures: bank of filters, fast Fourier transform (FFT), wavemeter, and swept spectrum or superheterodyne. Each has its own set of strengths and weaknesses, but in this article only the superheterodyne architecture will be discussed.

Figure 2 is a simple block diagram of a superheterodyne spectrum analyzer. It is essen-

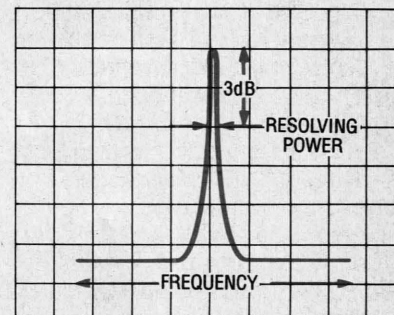


FIG. 3—TYPICAL CONTINUOUS-wave signal

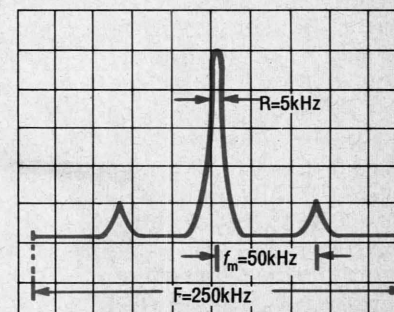


FIG. 4—SINGLE-TONE AMPLITUDE modulated signal.

tially a narrow-band superheterodyne receiver that is repeatedly swept in frequency over a selected portion of the RF band. Heterodyne means to mix signals together to produce sum and difference frequencies. Superheterodyne has the same meaning except that the signals being mixed are above the audio range.

The ramp (or sweep) generator generates a sawtooth (or ramp) waveform that tunes the local oscillator (LO). A form of frequency generator, it is swept linearly between two frequency limits. At the same time, the ramp generator also deflects the CRT beam horizontally across the screen from left to right. The local oscillator output is fed into the mixer.

The input signal is passed through a low-pass filter to the mixer, where it is mixed with the LO output. The output of the mixer includes not only the input and LO frequencies, but also their harmonics and sums and differences.

Only those mixed signals that fall within the passband of the intermediate frequency (IF) filter will be passed on to the envelope detector, represented here as a diode. The output of the envelope detector is amplified and applied to the vertical plates of the cathode-ray tube (CRT) to produce a vertical deflection on the CRT screen.

The CRT screen typically has a graticule or overlay grid of lines—ten horizontal and eight or ten major vertical divisions spaced one centimeter apart. The horizontal axis is calibrated in frequency that increases from left to right.

Signal treatment

If a spectrum analyzer receives a continuous-wave signal, the trace is a plot of the bandpass characteristic of the intermediate-frequency amplifier, as shown in Fig. 3. A single response will show up on the screen, and the resolving power of the analyzer is equal to the 3 dB bandwidth of the intermediate-frequency amplifier.

The spectrum of a single-tone, amplitude-modulated sig-

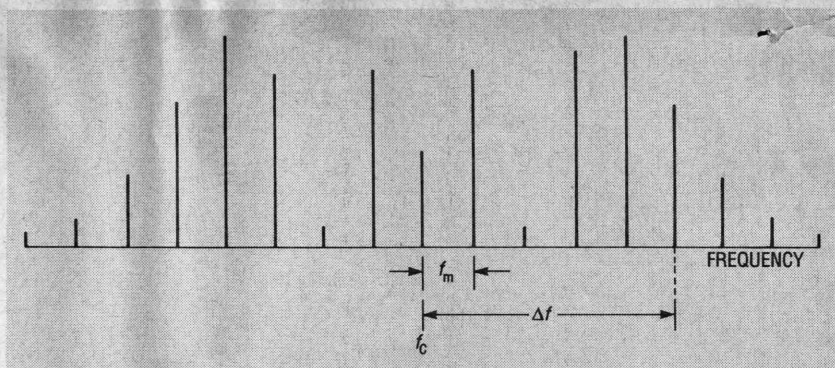


FIG. 5—SINGLE-TONE FREQUENCY-modulated tone where f_m equals the modulating frequency.

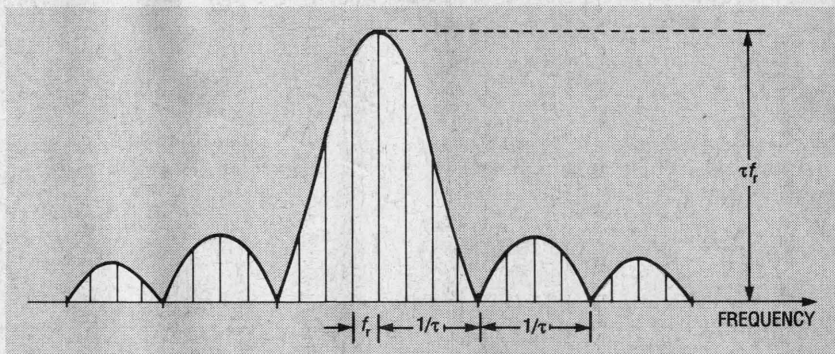


FIG. 6—PULSE-MODULATED SIGNAL where f_r equals the pulse repetition frequency and τ equals the pulse width.

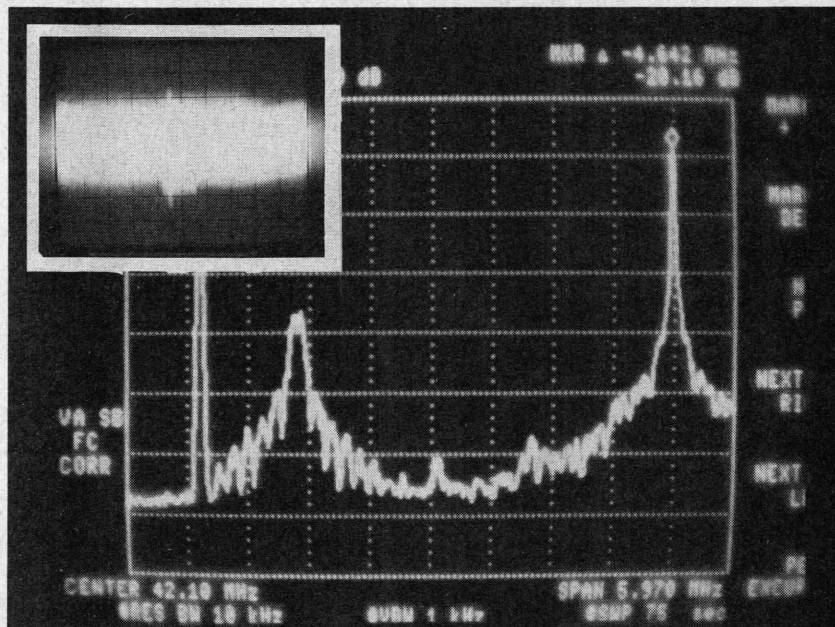


FIG. 7—TIME-DOMAIN AND FREQUENCY-DOMAIN PHOTOS: A time-domain waveform as displayed on an oscilloscope (inset), and frequency-domain waveform of the same signal as displayed on a spectrum analyzer.

nal will consist of the vertical response representing the original carrier frequency plus a pair of side frequencies, one above and the other below the carrier, as shown in Fig. 4. The amplitude of either sideband voltage with respect to the carrier volt-

age is $m/2$, where m is the percentage modulation. The frequency difference between the carrier and either sideband is equal to the modulating frequency, as shown in Fig. 4.

If the modulating wave is complex, each frequency com-

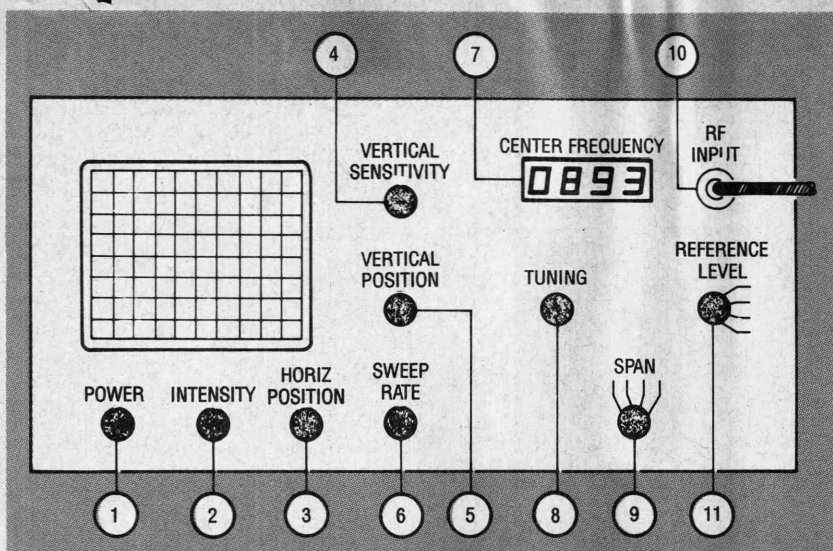
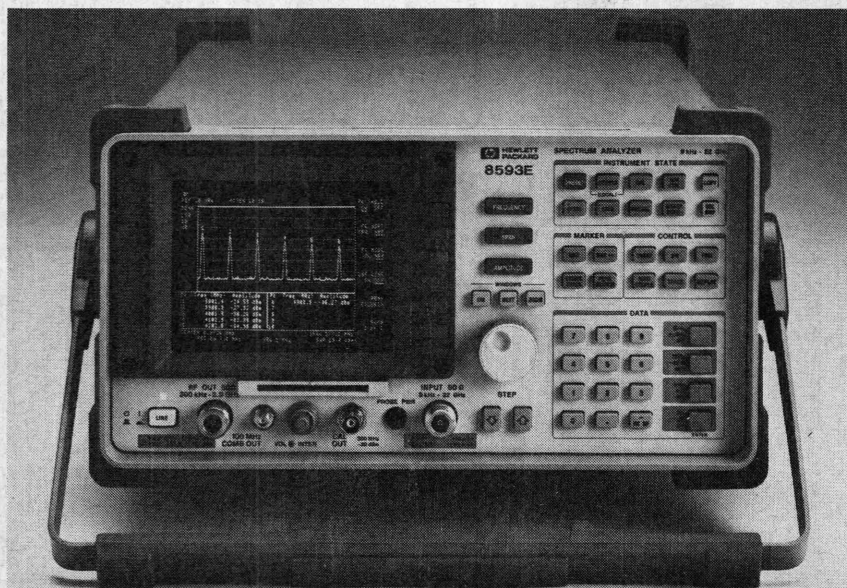


FIG. 8—SPECTRUM ANALYZER FRONT PANEL: The controls of a basic model.

1. POWER switch turns power on and off.
 2. INTENSITY controls brightness of the display.
 3. HORIZONTAL POSITION shifts the display on the screen.
 4. VERTICAL SENSITIVITY selects an amplitude sensitivity.
 5. VERTICAL POSITION moves the CRT display vertically.
 6. SWEEP RATE controls the sweep speed across the CRT
 7. CENTER FREQUENCY is a readout of center frequency.
 8. TUNING centers the frequency on the CRT display.
 9. SPAN controls the spectrum display width.
 10. RF INPUT accepts the signal to be observed.
 11. REFERENCE LEVEL sets waveform to the graticule top.
- More advanced models have additional function knobs, switches, and pushbuttons.



PORTABLE SPECTRUM ANALYZER from Hewlett-Packard covers the RF and microwave frequency range from 9 kHz to 22 GHz. It offers frequency-synthesized accuracy of 7.6 kHz at 1GHz. Second- and third-order dynamic ranges are 77 dB and 90 dB, respectively. Memory cards can quickly transform the unit into an application-specific instrument.

ponent in the wave contributes a pair of side frequencies to the spectrum. These side frequencies are located at equal distances on each side of the carrier, and they are spaced

away from the carrier by the modulating frequency of the wave under study.

In frequency modulation, a single modulating tone will produce a carrier with an infinite

number of side bands. Nevertheless, the spectrum will be symmetrical about the center frequency, as shown in Fig. 5.

Pulse modulation is a special case of amplitude modulation in which the carrier is on during selected intervals (pulses), and is off between these pulses. It is done to increase the ratio of peak to average power in the modulated wave as in pulsed radar or to encode a signal as in pulse code modulation (PCM).

Without going into the mathematical derivations, it is sufficient here to point out that the spectral display of a pulse-modulated carrier is that of a large magnitude *lobe* centered on the carrier's center frequency with side lobes of diminishing amplitude spaced equidistantly on either side of the center frequency, as shown in Fig. 6.

Graphic records

Figure 7 show a photograph of a 45.75 MHz, video-modulated, intermediate-frequency signal from a tuner, taken from an oscilloscope display. It can be seen that the modulation makes it impossible to freeze the display on this time-domain oscilloscope picture. Peak-to-peak voltage is the only measurement that can be accurately made.

Figure 7 also shows a spectrum analyzer presentation of the same waveform. It is now possible to recognize (reading from left to right) the audio subcarrier, the modulated 3.58 MHz chroma subcarrier, and the modulated video carrier with its sidebands. Because this is an IF output, the audio subcarrier is lower in frequency than the video signal.

From that waveform it is possible to measure the video carrier level, the audio carrier level below video, the frequency of the video carrier, and the frequency span from the video to the audio carriers. This is particularly important because an off-frequency audio subcarrier causes an annoying distortion on the TV screen called "beats."

With tedious adjustments and care, it might be possible to measure modulation depth, au-

dio deviation, and signal-to-noise ratio. However, none of these can be measured with a conventional oscilloscope.

Basic control functions

Figure 8 shows the front face of a spectrum analyzer with a minimum number of controls. The input connector is usually placed on the front panel. The power switch typically has a lamp nearby to indicate if the power is on or off.

The controls includes knobs for setting waveform intensity, horizontal position, vertical sensitivity, vertical position, and sweep rate. A tuning knob sets the center frequency, which is displayed on a liquid crystal display panel. SCAN WIDTHS or SPAN (MHz/div) sets the scan width and REFERENCE LEVEL (dBm) sets reference level. Some instruments have a knob for setting for BANDWIDTH (BW—normal or wide).

The vertical axis is calibrated in amplitude. Most analyzers offer a choice of linear scale calibrated in volts or power, or a logarithmic scale calibrated in decibels. The log scale offers a much wider usable range. The top line of the graticule is normally assigned as the reference level, and scaling per division permits the assignment of values to other graticule locations. This permits the measurement of both the absolute signal value or the amplitude difference between any two signals.

The cost of spectrum analyzers has been declining over the past decade while their performance, efficiency and reliability has been improved. The availability of higher levels of integrated circuitry has permitted lower power consumption.

The battery-portable 2610 spectrum analyzer from B+K Precision, shown on the first page of this article, covers the frequency range of 1 MHz to 1 GHz. It offers a fixed 1 MHz bandwidth, regardless of scan width. This, B+K says, simplifies the observation of video, television, and cable television signals. This B+K instrument is considered to be a general purpose analyzer.

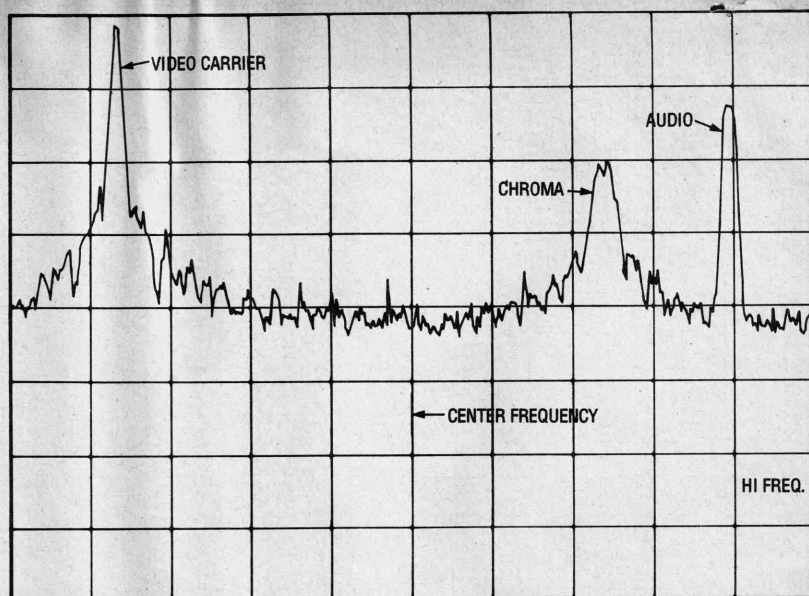


FIG. 9—THE STANDARD VIDEO CARRIER WAVEFORM (V) shown with its chroma (C), and audio (A) subcarriers at higher frequencies. Table 1 gives the basic settings.

TABLE 1
INITIAL SETTING ON SPECTRUM ANALYZER

Control	Abbrev.	Value	Units
1. Reference level	REF	9.3	dBmV
2. Attenuation		10.0	dB
3. Decibels/division		10.0	db/div.
4. Tune center freq.	CENTER	63.96	MHz
5. Residual bandwidth	RES BW	30.0	kHz
6. Visual bandwidth	VBW	30.0	kHz
7. Span	SPAN	6.0	MHz
8. Sweep rate	SWP	600	msec

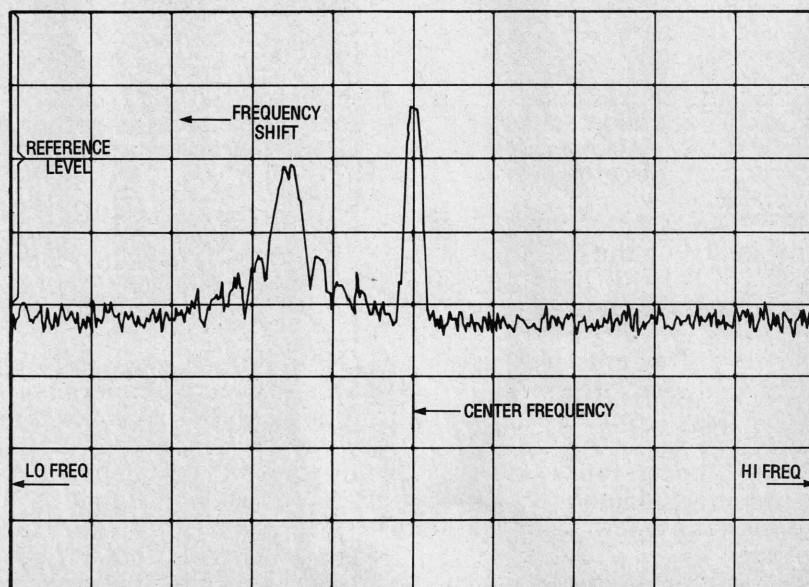


FIG. 10—RETUNING THE CENTER FREQUENCY from 63.96 to 66.33 MHz shifts the trace toward the higher frequency. All other settings remain the same.

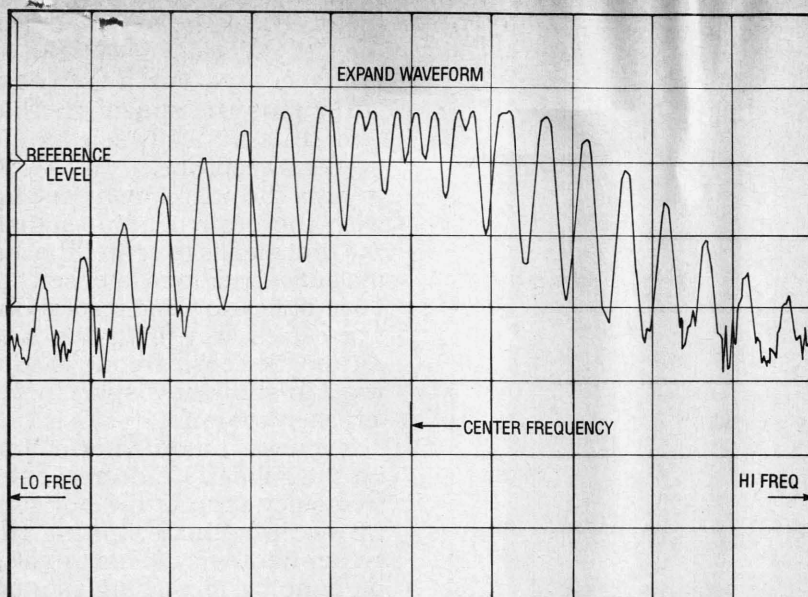


FIG. 11—DECREASING THE SPAN FROM 6.0 MHz to 200.0 kHz expands the trace. The center frequency has shifted from 66.33 to 66.31 MHz, and the sweep has decreased from 600 to 20 milliseconds. All other settings remain the same.

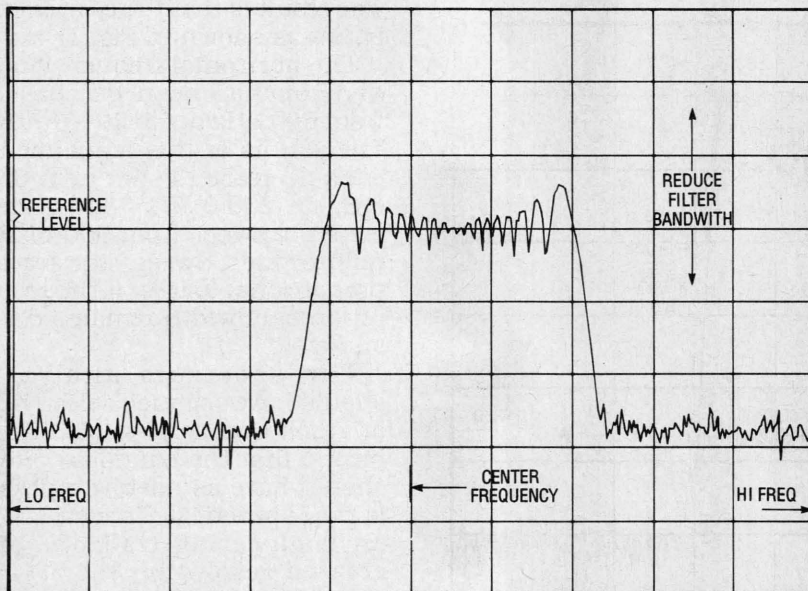


FIG. 12—DECREASING THE RESOLUTION BANDWIDTH from 30 kHz to 1 kHz shows the true amplitude of the sidebands. The sweep has increased from 20 milliseconds to 18 seconds. All other settings remain the same.

The high end of the market is represented by the 8593E from Hewlett-Packard. It is a spectrum analyzer that can make frequency measurements at 22 GHz. One of a family of HP spectrum analyzers, it is designed so that relatively unskilled operators can make routine specialized precision measurements on the factory floor just by pushing a few front-panel buttons.

Some of the high-end models include microprocessors which permit the display of alpha-

numeric legends on the CRT screen to verify settings. It also permits the selection of amplitude units on either log or linear scale.

Some models have movable cursors that can be located at any point on the displayed trace so that the frequency or amplitude values at that point will be displayed. Digital storage of the displayed waveform eliminates display flicker at the sweep rate. Pushbuttons have replaced control knobs on some models, and battery-portable analyzers are

also becoming popular.

Advanced spectrum analyzers have provision for "personality cards," which, when inserted, dedicate the instrument to specific applications and minimize the number of settings and adjustments that must be made. This feature permits relatively unskilled operators to carry out precise production-line or laboratory measurements.

Control settings

The two basic settings that must be made when operating an oscilloscope are vertical volts per division and horizontal time (seconds) per division. Of course, other settings such as trigger level, horizontal and vertical position, and certain other enhancements will give the clearest possible waveform.

By contrast, when operating a spectrum analyzer, there are more required settings. These include the vertical reference level, horizontal center frequency, horizontal sweep time, and resolution bandwidth.

Some analyzers offer a choice of vertical reference level units. These include decibels referred to a milliwatt (dBm), to a nanovolt (dBnV), to a microvolt (dBμV), and to volts and watts. A choice can also be made between a 50-ohm and a 75-ohm input impedance.

Figure 9 is a rendering of an actual spectrum analyzer waveform of a radio-frequency video carrier reproduced with an X-Y plotter. Notice that because it is a radio-frequency signal and not an intermediate-frequency video signal, the carriers are in their normal positions—the audio has a higher frequency than the video signal. Table 1 gives the initial settings.

Referring back to the spectrum analyzer front-panel diagram, Fig. 8, the following information can be gained:

- **Reference level**—The reference level is given at the top of the graticule. All other voltage or power levels are below that maximum level. The REFERENCE LEVEL knob adjusts the input attenuator and IF gain so that the top graticule corresponds to the

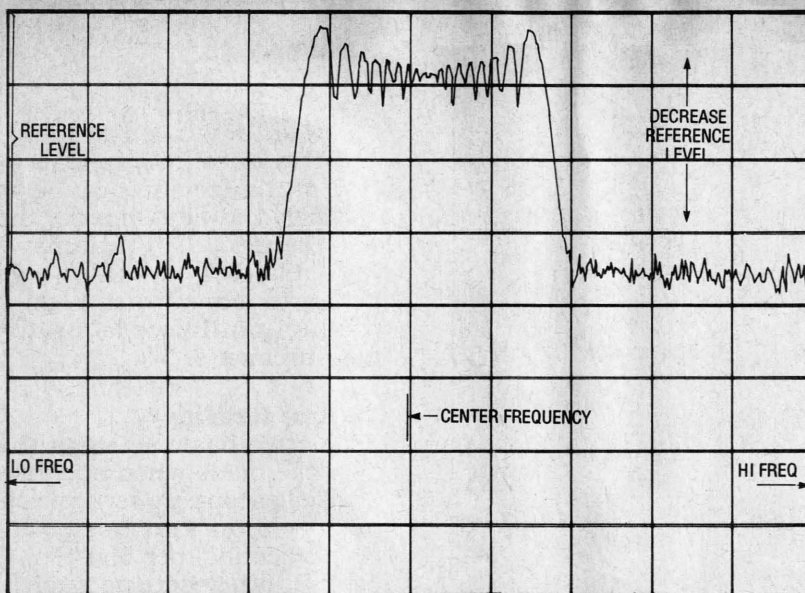


FIG. 13—DECREASING THE REFERENCE LEVEL from 9.3 dBmV to -12 dBmV elevates the waveform. All other settings remain the same.

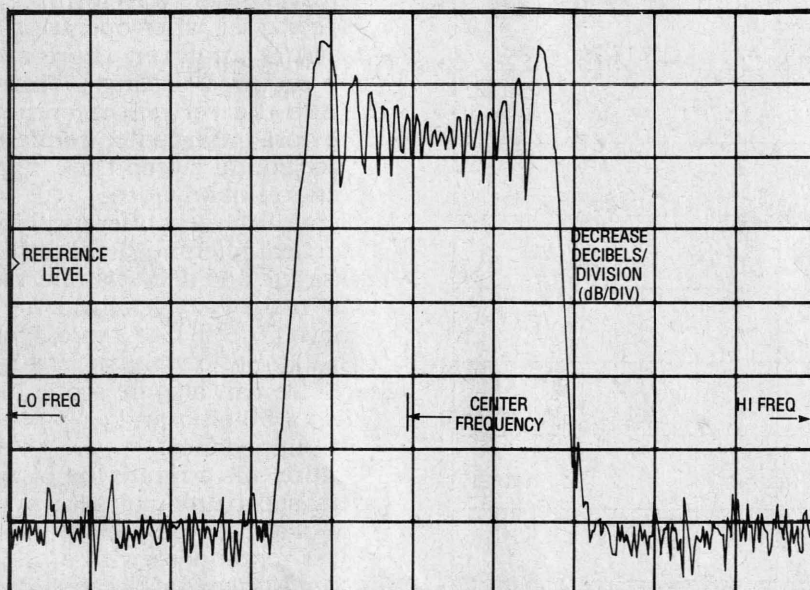


FIG. 14—DECREASING THE DECIBELS/DIVISION value from 10 to 5 dB/div amplifies the waveform. All other settings remain the same.

indicated signal level. Calibrations might be in dBm or dBmV.

- **Vertical sensitivity or scale**—selects the amplitude sensitivity of the graticule. Typical selections are 10 dB/div or 2 dB/div.

- **Tuning**—sets the mean or center frequency of the display. Some models display this value separately on a liquid-crystal display.

- **Span**—controls the width of the spectrum being displayed, and automatically selects a filter for optimum resolution. It is the

difference between the low and high frequencies.

- **Low frequency**—A value that can be determined by subtracting half the value of the span from the center frequency.

- **High frequency**—A value that can be determined by adding half the value of the span to the center frequency.

- **Resolution bandwidth: RES BW**—A setting that is unrelated to signal amplification, but it affects the resolution of the waveform presentation.

- **Sweep rate: SWP**—controls

the speed of the sweep across the CRT, typically measured in milliseconds. For most measurements the spectrum analyzer automatically selects the proper sweep time.

With the instrument set up with the desired input signal, the first step is to center the audio subcarrier for a later setting of amplification and resolution. The easiest way to do this is to retune the center frequency upward to shift the display to the left as shown in Fig. 10.

If you want more information on the subcarrier, decrease the frequency span of the horizontal sweep without shifting the center frequency. (This can also be done by increasing the low frequency and decreasing the high frequency the same amount in the same direction.) The result is that the carrier expands as shown in Fig. 11.

The horizontal frequency per division has been decreased from 600 kHz/div to 20 kHz/div. The results of this adjustment are a decrease in SPAN from 6.0 MHz to 200.0 kHz, and a decrease of SWEEP from 600 to 20 milliseconds. Sweep time SWP is proportional to SPAN if the resolution bandwidth remains constant.

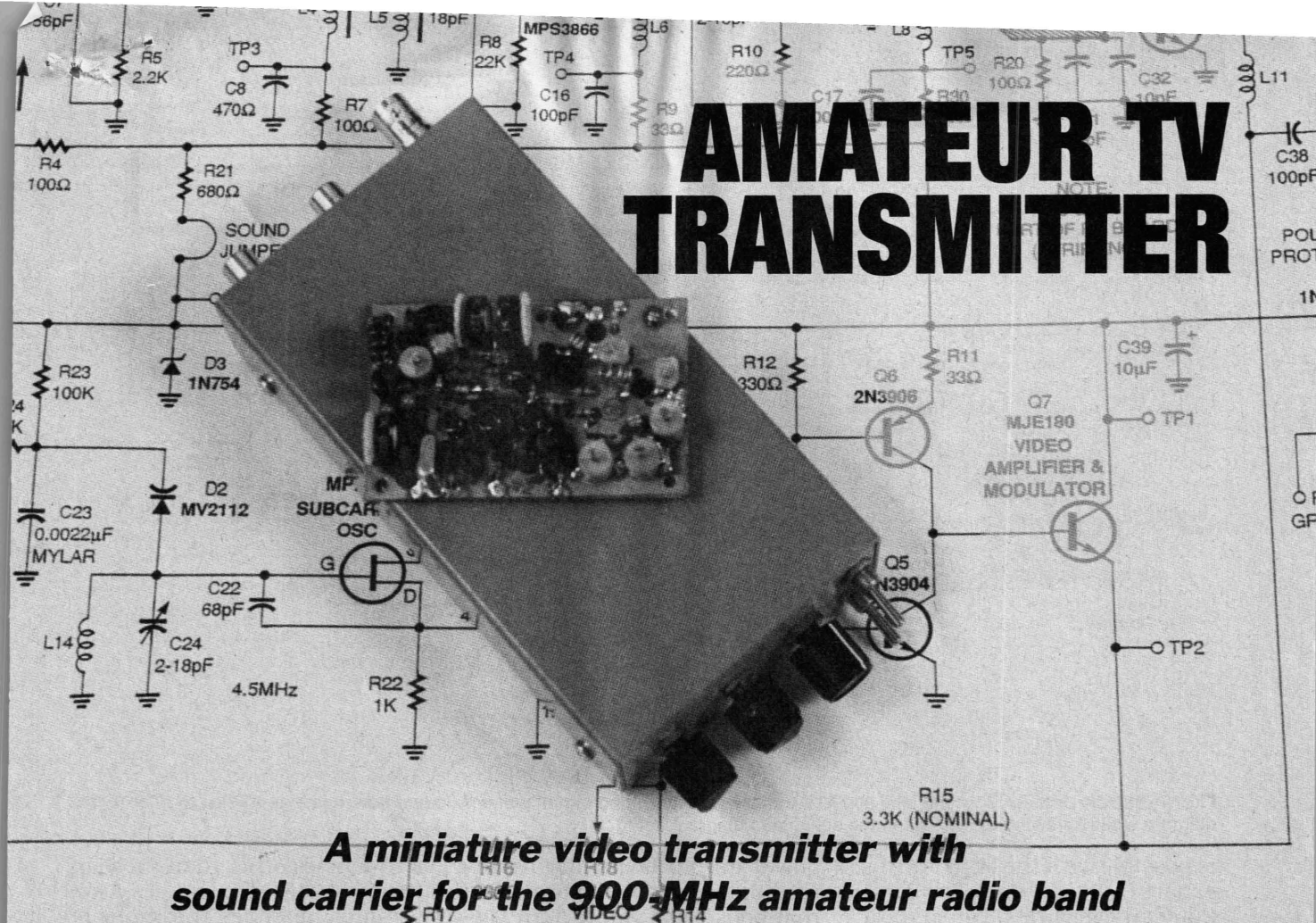
The spectrum analyzer should have as much selectivity or resolution as possible. This means that the bandpass filter should have as narrow a pass-band as practical. This calls for an engineering tradeoff: increased resolution increases sweep time (SWP) because sweep time is inversely proportional to the square of the resolution bandwidth.

Decreasing resolution bandwidth by half increases sweep time by a factor of four. There is a practical limit because sweep time would become excessive, measurable in tens of seconds rather than milliseconds. Fortunately, it is rare that there is ever a need for resolution so high that those unusual sweep times would be encountered. The rule is to keep sweep time as fast as possible.

In the example given here, there is a need for increased res-

Continued on page 86

AMATEUR TV TRANSMITTER



A miniature video transmitter with sound carrier for the 900-MHz amateur radio band

William Sheets, K2MQJ & Rudolf F. Graf, KA2CWL

YOU CAN BUILD A MINIATURE AMATEUR radio TV transmitter and put it to work in a variety of applications from radio-controlled videography to two-way communications—and just about any other video application that might strike the fancy of experimenters and amateur radio operators. Now, small CCD cameras help to keep the cost low.

Don't confuse the Mini 900 ATV transmitter with video-link devices operating in the 900-MHz band. The Mini 900 is very small ($1.75 \times 2.75 \times 0.75$ inch), develops 500 milliwatts of RF output, is crystal controlled, and has produced clean pictures at distances over 5 miles, using simple antennas at both ends of the transmission path. Note that the 902-928 MHz band is available to ham radio operators holding a valid Technician Class license or higher. A license is required to operate this project in the U.S. (No Morse code test is required

for the Technician license.) The transmitter is for experimental non-commercial use only and its application is subject to radio amateur rules and regulations in the U.S. and other countries.

Operation of the Mini 900 ATV transmitter on commercial TV frequencies is illegal in the U.S. and for other nations. The device has *not* been submitted for FCC type acceptance and is *not* legal for commercial use.

The Mini 900 must be operated so that the entire RF signal lies within the 902-928-MHz band when operated in the U.S. This means picture carrier frequencies of 906.5 and 923.5 MHz inclusive. A good choice for most radio amateurs is 910.25 MHz. U.S. operators should check their local areas first. Other common carrier frequency alternates are 922.25 and 923.25 MHz.

The Mini 900 accepts an NTSC video signal and gener-

ates a video-modulated RF signal including a standard 4.5-MHz sound subcarrier in the 902 to 928-MHz band. The signal from the Mini 900 can be received on any standard TV receiver fitted with a suitable RF downconverter capable of tuning the 902-928 MHz Amateur band. The transmitter's small size (1.75×2.75 inches) is comparable to many small PC board cameras and is small enough for radio-controlled applications in which a video link is needed.

The Mini 900 can operate from a 10 to 14-volt DC power supply, with minor adjustments in the transmitter's video drive. Lead acid, NiCd, or alkaline power packs may be used. RF power output will be 0.2 to 0.7 watts over the power-supply voltage range. Nevertheless, the author used a 13.2-volt DC power supply and it is referenced throughout this article.

Operation over 900-940 MHz

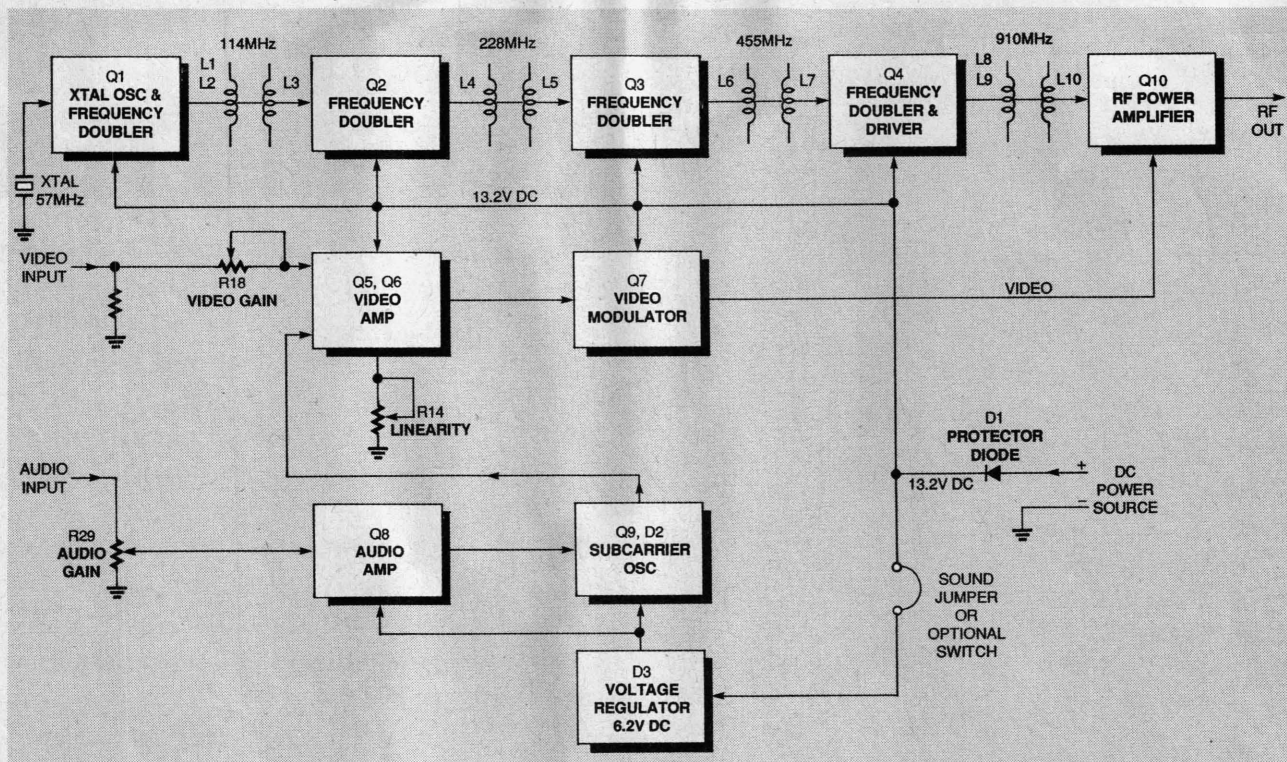


FIG. 1—BLOCK DIAGRAM for the Mini 900 ATV transmitter illustrating signal generation and flow. It is a good idea to refer to this diagram while following the details in the schematic diagram in Fig. 2.

is possible with the Mini 900 for export (non-US) use. The Mini 900 can easily be retuned for a 5.5 MHz sound subcarrier used by the PAL system. The video input requirement is standard 1-volt peak-to-peak, negative sync signal. The video amplifier and modulator are DC coupled, but a capacitor can be installed in series with the input if AC coupling is desired. Audio inputs from 5 millivolts to 1 volt can be accommodated.

This project can be built from scratch, in two to four evenings, depending on the builder's experience. Due to the unit's small size and many small parts, and the high frequencies present, this project is not recommended for beginners.

How it works

The Mini 900 consists of a crystal oscillator operating around 57.5 MHz, a series of frequency multiplier stages to multiply the crystal oscillator frequency by a factor of sixteen, and an RF power amplifier. Refer to Fig. 1 for the functional block diagram of the Mini 900 circuit and Fig. 2 for a detailed schematic diagram. The RF-

power amplifier is amplitude-modulated by a video amplifier that boosts the video input signal to a required level. An audio amplifier stage boosts the audio input which in turn frequency modulates a 4.5 MHz VCO (voltage-controlled oscillator) to generate a sound subcarrier. The FM subcarrier is combined with the video in the modulator. If no audio is needed this feature can be disabled by disconnecting a jumper.

Transistor Q1 is a crystal-controlled oscillator using a 56–59 MHz crystal, XTAL1, in a common-base transistor circuit (see Fig. 2). At the series-resonant frequency of the crystal, the base of Q1 is nearly grounded. Coils L1 and C2 are tuned slightly above resonance, causing oscillation at the crystal frequency. The signal generated is rich in harmonic frequencies.

A filter, consisting of L2, C3, C4, L3, C6 and C7, couples the second harmonic of the crystal's 112-118 MHz signal to frequency-doubler stage Q2. This stage further doubles the signal frequency to 225-240 MHz. A filter, consisting of C9, L4, C10, L5, C11, and C37, suppresses all

frequencies other than those in the 225-240 MHz range. Also, the filter matches the input of the next stage, frequency doubler Q3. Transistor Q3 doubles this signal to half the final output frequency, 450 to 470 MHz. At this point, about 30 milliwatts of RF power is present.

A filter, consisting of C12, L6, C15, C13, C14, and L7, forms a matching network to couple the RF energy from Q3 to Q4, and also filters out unwanted signals. Transistor Q4 doubles this signal to the final output frequency, 900-940 MHz, supplying about 50 to 80 milliwatts of RF driving power. A double-tuned coupling network, consisting of C18, C19, L9, L10, C20, C31, and C32, matches the output of Q4 to the base of Q10. This network has tuned strip-line coupled inductors (see Fig. 2) that are actually copper strips etched on the PC board.

RF power amplifier Q10 supplies an RF output of between 0.2 and 0.7 watt depending on supply voltage. A coupling network, consisting of L11, C33, L12, C34, C35, L13, and C36, matches a 50-ohm antenna load

Continued on page 61

ATV TRANSMITTER

continued from page 32

connected to the collector of Q10 and suppresses harmonic frequencies. The RF output is a relatively pure signal between 900-940 MHz.

Video and audio modulation

To apply video modulation to this RF carrier, the supply voltage for Q10 is taken from the emitter of video modulator Q7. Transistors Q5, Q6, and Q7 are connected as a feedback amplifier with a gain of about 5 to 10. Video input to the transmitter is amplified in this circuit and also inverted in polarity so that sync and black levels are positive going. The video signal at the emitter of Q7 is superimposed on the supply voltage to Q10. In turn, Q10 amplitude modulates the RF output from Q4. Potentiometer R14 sets the carrier level of the RF output for symmetrical modulation and

potentiometer R18 sets the modulation level for maximum without distortion or clipping.

The Mini 900 also provides a 4.5-MHz audio subcarrier. Incoming audio is fed to gain control R29. Audio from R29 then is fed through C27 to the base of Q8. Audio from Q8 is fed to varactor diode D2 via C25 and R24. Diode D2 is part of the frequency-determining network of 4.5-MHz subcarrier oscillator Q9. Circuit elements D2, C24, C21, C22, and L14 make up the 4.5-MHz tuned circuit. Audio voltage on D2 changes its capacitance, causing frequency modulation to occur. Trimmer capacitor C24 sets the 4.5 MHz sound subcarrier frequency.

The subcarrier output from Q9 is fed into the video modulator via C30, where it is mixed with the video. A jumper in series with the supply to 6.2-volt Zener diode D3 and R21 permits disabling Q8 and Q9 if no sound is necessary. An optional external switch or removed

jumper will power down the sound circuit for longer battery life. Diode D3 and C29 supply a clean, regulated 6.2 volts bias to Q8 and Q9. Resistor R21 and C28 decouple the sound circuit from the DC supply. For PAL systems the number of turns on L14 is reduced to allow 5.5-MHz operation.

Diode D1 protects the Mini 900 circuitry against accidental reverse polarity from the external power supply and C39 provides supply line bypassing.

The DC voltage supply to the transmitter should be very clean and stable. Any noise or ripple in the supply voltage can modulate the RF carrier and cause severe distortion in the received video. Less than 50 millivolts of ripple and noise is permissible, less being more desirable. Dedicated battery packs offer the best option for a noise-free voltage source, otherwise voltage regulation and additional filtering should be considered.

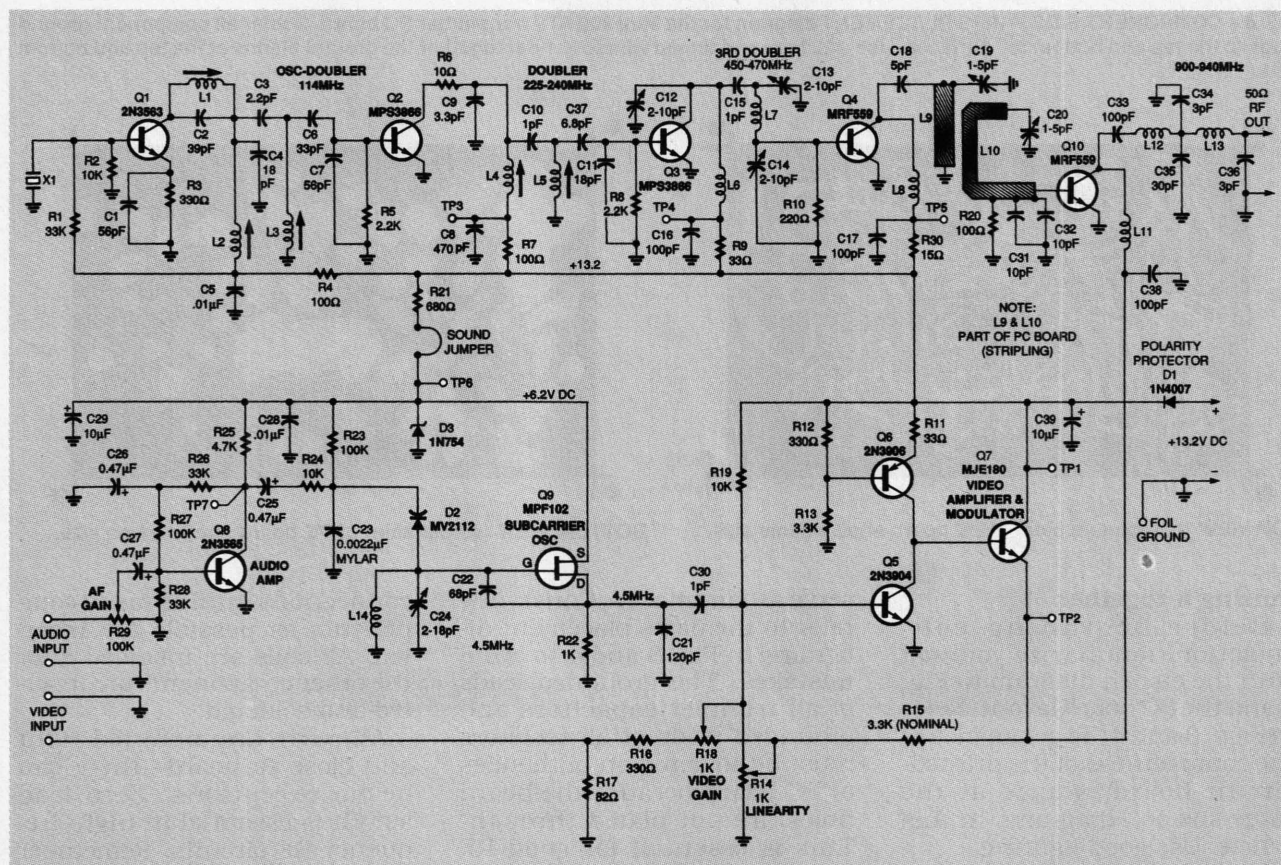


FIG. 2—SCHEMATIC DIAGRAM for the Mini 900 ATV transmitter. All the parts in the diagram fit on a 2 3/4 × 1 3/4-inch two-sided foil printed circuit board. An understanding of the circuit diagram will enable builders to easily perform alignment and troubleshooting procedures.

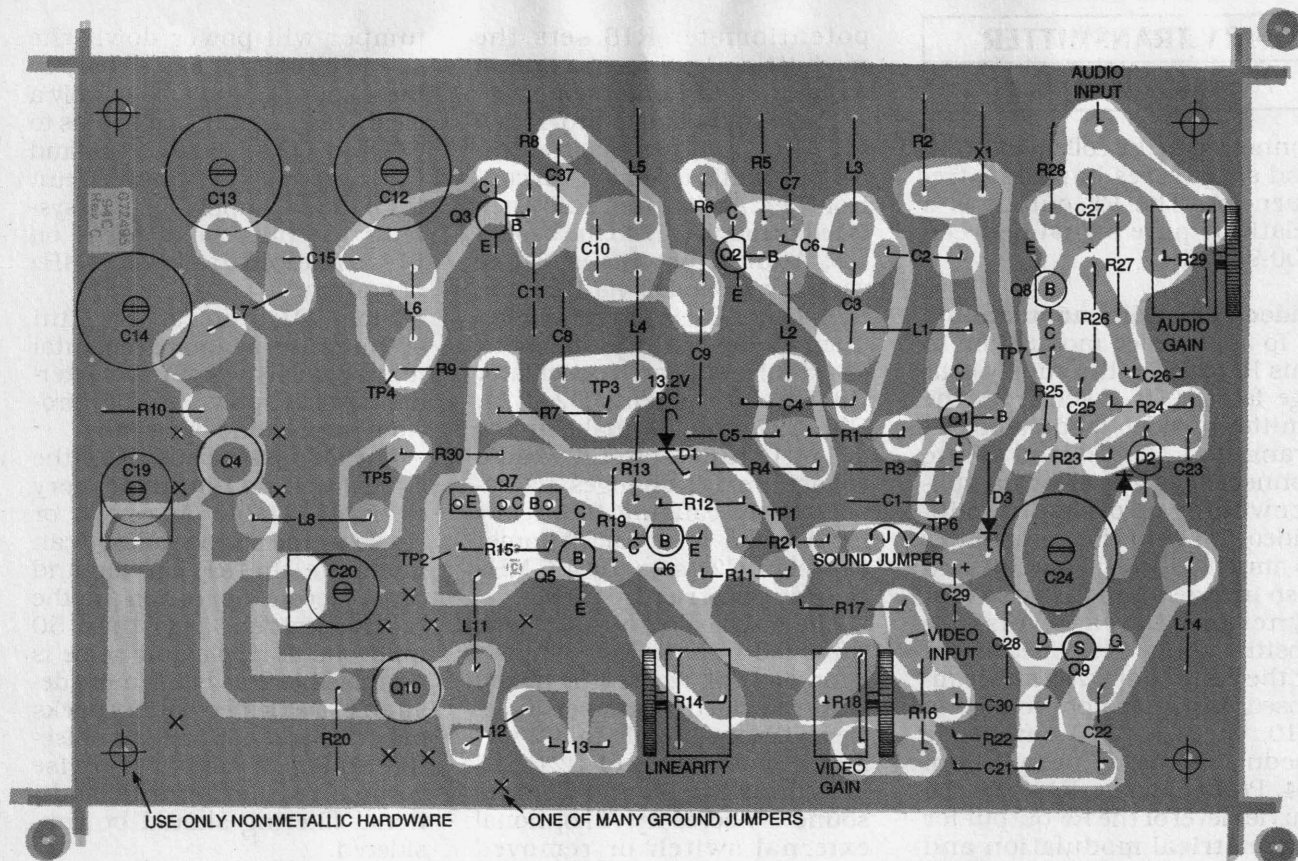
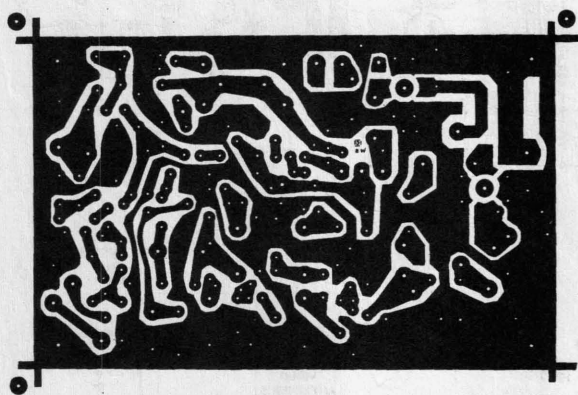
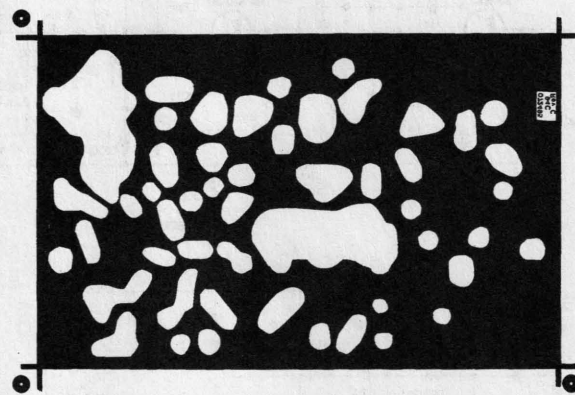


FIG. 3—COMPONENT SIDE PARTS PLACEMENT diagram for the Mini 900 ATV transmitter PC board. Solder all component ground leads to the top and bottom foil surfaces. Use #22 bare and tinned wire to jumper together the ground planes of the top and bottom foils.



TOP VIEW (component side) of PC board shown same size.



BOTTOM VIEW (solder side) of PC board shown same size.

Putting it together

Before beginning construction, familiarize yourself with the circuit diagram in Fig. 2 and the PC board layout. Refer to Figs. 3 and 4. To get an idea of the compactness of the printed-circuit board, glance at the same-size foil diagrams in this article. Use good lighting.

It is helpful to insert first a few large parts such as trimmer capacitors and potentiometers to

serve as landmarks. Constantly refer to the parts placement diagrams in Fig. 3 and 4 to avoid mistakes. The grounded leads of all trimmer capacitors and grounded leads of all resistors must be soldered on both sides of PC board because the board holes are not plated through. This is essential for good RF grounding. Except where access may be difficult later in construction, *do not* solder any

connections until as many components as possible are inserted. All coils are installed after the other components are inserted and soldered.

All parts are mounted tight and close to board—there can be no exceptions. *Zero* lead length is essential in high-frequency RF circuits. Remember that the Mini 900 operates at UHF frequencies and lead lengths are critical—they intro-

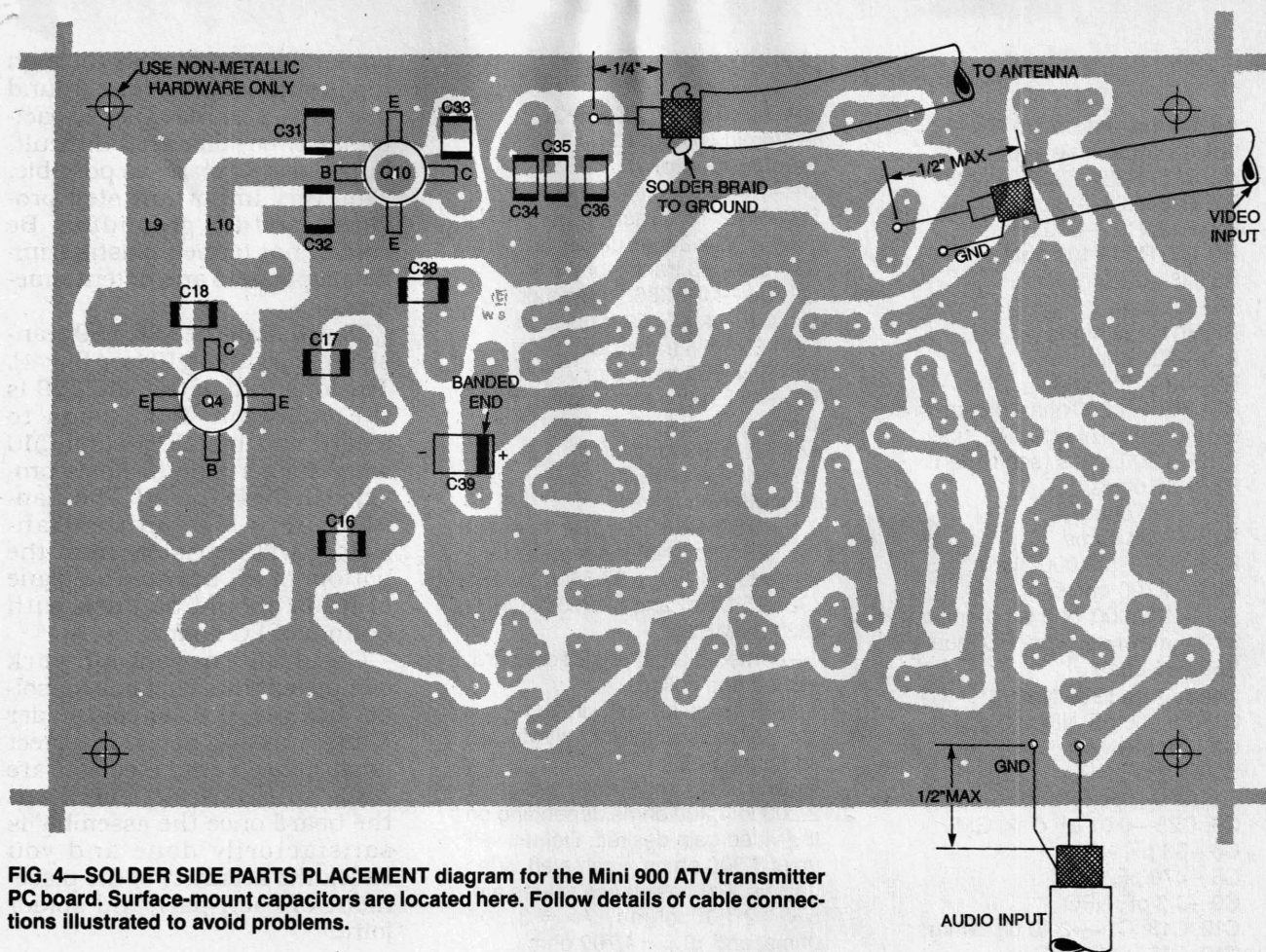


FIG. 4—SOLDER SIDE PARTS PLACEMENT diagram for the Mini 900 ATV transmitter PC board. Surface-mount capacitors are located here. Follow details of cable connections illustrated to avoid problems.

duce unwanted inductance and stray capacitance.

Install all fixed resistors and solder in place (see Fig. 3). The next step is to install and solder trimmer side terminals to top side of the PC board. Be careful not to melt the plastic. Do not solder the bottom at this time. Install all ground-plane jumpers that connect top and bottom ground planes (See Fig. 3).

Install capacitors with wire leads. Pre-formed leads should be straightened with pliers so as to fit as close to board as possible. Watch polarity of all electrolytic capacitors so as not to accidentally reverse polarity. Do not install surface-mount capacitors until a later step. Install all transistors, except Q4 and Q10. Check for correct orientation of transistor leads. All transistors should be 1/8 inch from surface of board. Solder the emitter leads of Q2 and Q3 to the top of the PC board where they pass through the ground foil.

Install D1, D2 and D3, checking the polarity (marking band) before and after soldering. Solder the ground lead of D3 to the top side of the PC board where it contacts it. Note that D1 (1N4007) has only one lead inserted in board (banded end). The other end is formed into a hook and is used as a terminal for attaching the DC supply positive lead wire. Diode D2 looks like a transistor with only two leads. Refer to Fig. 3 before installing D3.

Using scissors, trim the crystal leads to $\frac{3}{16}$ inch and install. Do not use diagonal cutters as the mechanical shock produced during cutting can damage the crystal. Solder a short wire to the top of the crystal and run this lead to ground.

Carefully fabricate all coils and install them in the PC board. The winding of the coils are detailed in a sidebar entitled "Coil Winding Data." Be sure that no turns short together on L6, L7, L8, L11. The bottom

turn of L1 connects to the collector of Q1: L2 bottom turn, to the junction of C3 and C4; and L3 bottom turn, to the junction of C3 and C6. Use an 8-32 screw to hold L1, L2, L3, L4 and L5 during installation. After installing coils, remove screw and insert tuning slugs fully into coils. Note that coils L11, L12, L13 are fabricated on a 6-32 screw while all others except toroid L14 use an 8-32 screw as a winding form. Use a small drop of paraffin wax or candle wax to retain tuning slugs in coils.

Install C16, C17, C33, C38 (100-pF surface-mount capacitors); C18 (5-pF surface-mount capacitor); then C31 and C32 (10-pF surface-mount capacitors); C34, C35 and C36 (3-pF surface-mount capacitors) and C39 on the underside of the PC board. Refer to Fig. 4 for capacitor location and sidebar entitled "How to Install Surface-mount Capacitors" for soldering details. Caution: Check the polarity on C39 as it is soldered in

PARTS LIST

All fixed resistors are 1/8-watt, 5%, unless otherwise specified.

R1, R26, R28—33,000 ohms
R2, R19, R24—10,000 ohms
R3, R12, R16—330 ohms
R4, R7, R20—100 ohms
R5, R8—2,200 ohms
R6—10 ohms
R9, R11—33 ohms
R10—220 ohms
R13—3,300 ohms
R14, R18—1,000 ohms potentiometer, thumbwheel type, PC mount
R15—3,300 ohms (see Note 1)
R17—82 ohms
R21—680 ohms
R22—1000 ohms
R23, R27—100,000 ohms
R25—4,700 ohms
R29—100,000 ohms potentiometer, thumbwheel type, PC mount
R30—15 ohms

Capacitors (See Note 3)

C1, C7—56 pF, NPO
C2—39 pF, NPO
C3—2.2 pF, NPO
C4, C11—18 pF, NPO
C5, C28—0.01 μ F disk, GMV
C6—33 pF, NPO
C8—470 pF, GMV
C9—3.3 pF, NPO
C12, C13, C14—2-10 pF trimmer
C10, C15—1 pF, NPO
C16, C17, C33, C38—100 pF, surface-mount type
C18—4.7 or 5.0 pF, NPO, surface-mount (use either value)
C19, C20—1.5-5 pF trimmer
C21—120 pF, NPO or SM
C22—68 pF, NPO
C23—0.0022 μ F, 50-volt, Mylar
C24—2-18 pF trimmer
C25, C26, C27—0.47 or 1.0- μ F, 35-volt tantalum electrolytic (use either value)
C29—10 μ F, 16-volt electrolytic
C30—1 pF or 2.2 pF, NPO (see Note 2)
C31, C32—10 pF, NPO surface-mount
C34, C35—3 pF, NPO
C36—30 pF, NPO
C37—6.8 pF, NPO
C39—10 μ F, 16-volt, surface-mount

Inductors

See Coil Wind Chart for details of coil construction.

L1, L2, L3, L4, L5, L14—No. 22 enameled solid-copper wire
L6, L7, L8, L11, L12, L13—No. 22 bare tinned solid-copper wire

5—Ferrite slugs, blue, 8-32

1—Toroid core for L14

Semiconductors

D1—1N4007 rectifier
D2—MV2112 varactor
D3—1N754 Zener diode, 6.2-volts
Q1—2N3563 transistor
Q2, Q3—MPS3866 transistor
Q4, Q10—MRF559 transistor
Q5—2N3904 transistor
Q6—2N3906 transistor
Q7—MJE180 video amplifier modulator
Q8—2N3565 transistor
Q9—MPF102 JFET

Miscellaneous

X1—56.890625-MHz crystal (910.25 MHz). Note: Alternate crystal frequencies are 57.640625 MHz (922.25 MHz) and 57.703125 MHz (923.25 MHz).

1—printed-circuit, 2-sided board, etched and drilled.

1-8-32 screw for coil winding

1-6-32 brass screw for coil winding and tuning aid

Note 1: R15 resistance range is 2,200 to 4,700 ohms depending on the video gain desired. Default value is 3,300 ohms. Estimated video gain vs. R15 resistance values are: $5 \times = 2,200$ ohms, $7 \times = 3,300$ ohms, and $10 \times = 4,700$ ohms.

Note 2: Use either 1 pF or 2.2 pF for C30. Value selected determines sound subcarrier level.

Note 3: All capacitors less than 120 pF are NPO types with ceramic bodies. All capacitors with a specified working voltage are electrolytic types. All 0.01 μ F and 470 pF capacitors are GMV ceramic disk types.

A complete parts kit for the Mini 900 ATV transmitter, consisting of the PC board and all parts that mount on it, is available from: North Country Radio, PO Box 53, Wykagyl Station, New Rochelle, NY 10804. Kit is furnished with crystal for 910.25 MHz as standard, or alternately 922.25 or 923.25 MHz if specified with order. Kit price: \$81.50 plus \$4.50 for postage and handling. New York residents must add 8.25% sales tax. A catalog packed with cameras and other amateur TV kits and electronic projects is available. Please send SASE (75 cents) plus \$1.00 to cover handling (refundable with order).

ponent leads that pass through top ground and bottom ground planes as possible. Disc capacitors may be somewhat difficult, but solder as many as possible. This very important step provides good RF grounding. Be careful not to melt plastic trimmer capacitors and potentiometers.

Now install the MRF559 transistors, Q4 and Q10 (see Fig. 4). The long lead of the MRF559 is the collector. Trim leads to length and install Q4 and Q10 with thicker side in holes provided in the PC board. The manufacturer's logo on the transistors will be visible from the bottom of PC board. The plane of leads should be flush with bottom of PC board.

Carefully inspect all work completed thus far. Look for solder bridges, poor or cold solder joints, missing parts, incorrect parts placement, etc. You are ready to electrically check out the board once the assembly is satisfactorily done and you check the accuracy of the placement of each part and solder joint.

Checkout and tune-up

Do not proceed with this section until you have thoroughly checked out the construction as described in the last step of the instructions. You should have access to the suggested basic test equipment if you are to obtain optimum performance. It is futile to try to tune for best picture. It is, for example, possible to get a good picture and have only ten foot range of transmission. It is also possible to inadvertently overload the video monitor and get a lousy picture on it from a perfectly operating circuit board, and spend hours tracing nonexistent problems. Therefore use our procedure and experiment later when you are sure everything works properly.

The use of a metal standoff or other metallic material in the vicinity of L9 near the corner mounting hole in the PC board (See Figs. 3 and 4) can upset the stripline coupling network L9-L10. This will affect the setting of C19 and also may reduce the

place and afterwards: the banded side is positive (+).

Install sound jumper (J) near

D3 (see Fig. 3). Install all ground jumpers in the holes indicated in layout. Solder as many com-

attainable RF power output from the transmitter. For this reason, note the following precautions:

- Only use nylon or other non-metallic hardware in the corner hole next to coil L9.
- Keep the PC board at least ¼ inch from metallic or RF lossy materials.
- Keep any external wiring away from PC board especially near L9 and L10.

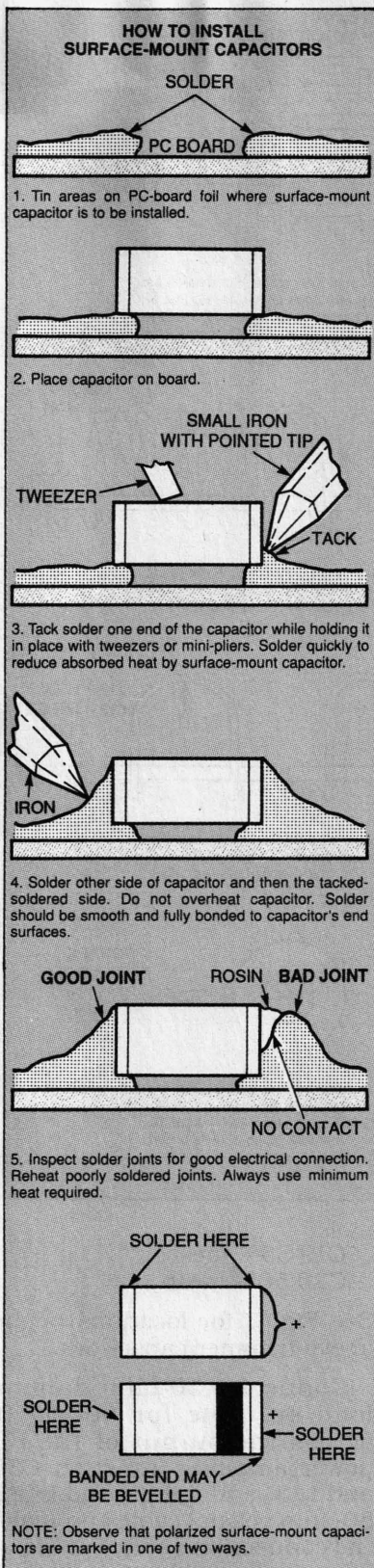
First verify that the DC voltages are correct and that the video modulator is working. Set the power supply to zero volts. Connect the negative lead of your power supply to the ground foil on the board. Next connect a lead from the positive terminal of the power supply through an ammeter (0-500 mA range) to the power input terminal which is the hooked end of diode D1. If the power supply has an ammeter, you can use it instead of an external ammeter. Now slowly raise the voltage to 10 volts. If more than about 100 mA is drawn, turn off the power supply and check your board for wiring errors, short circuits, solder bridges, etc. Power up again and if more than 100 mA is drawn, check to see if any part gets hot or smokes. If so, find the cause and repair assembly errors and/or damaged parts before proceeding.

Test points

Seven test points are shown on the schematic diagram in Fig. 2. These are intended for tune-up and diagnostic use on a finished, operating transmitter. Not all of the test points (listed below) are used in the tuneup procedure.

Test Point	Location	Volts DC
TP1	Collector, Q7	+13.2
-	Collector, Q1	+12.6
-	Collector, Q2	+13.2
TP6	Drain, Q9	+6.8
TP7	Collector, Q8	+3.5
-	Collector, Q4	>+2.0 to <+11.5*
TP2	Emitter, Q7	>+2.0 to <+11.5*

*These voltages depend on the setting of R14 and R18. Verify this before proceeding further.



If everything is satisfactory at 10 volts, set the power supply voltage to 13.2 volts. Connect a voltmeter negative lead to the power supply negative terminal

and check for following voltages using the positive lead of the voltmeter:

Set potentiometer R14 so that voltage at emitter of Q7 (TP1) is 6.0-volt DC. Connect a frequency counter between TP2 and ground. You should see a stable reading between 4 to 5 MHz. Set C24 for a counter reading of 4.495 to 4.505 MHz. This sets the sound subcarrier frequency. If you cannot obtain this frequency, but obtain a stable reading that is either too low or too high, remove or add a turn, respectively, to toroidal coil L14. You must have a *stable* reading. Ignore anything else. Use an oscilloscope to verify high-frequency oscillation at the source terminal of Q9. (For PAL applications this frequency will be 5.5 MHz and not 4.5 MHz.)

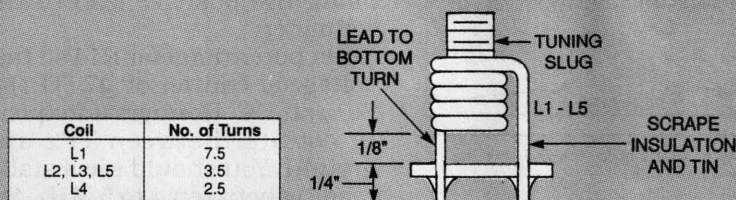
If all voltage tests are satisfactory, set potentiometer R18 at mid-range and set potentiometer R14 for maximum voltage reading on the emitter of Q7 (TP2). There should be at least 11.0-volts DC at TP2. If necessary, reset R18 so you get the 11 volts or more.

Tuning the RF circuits

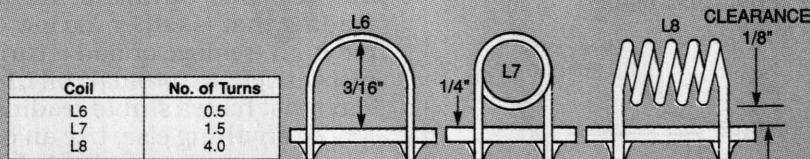
RF stages Q2, Q3, and Q4 and video amplifier Q5 are operated in class C mode. In the absence of an input signal, they are "off" and draw no current. An input signal causes the stages to turn on and draw a current that is somewhat proportional to the input signal. Therefore, by monitoring the collector current of a stage, the drive signal supplied to that stage can be monitored. Test points are used to monitor stage currents by reading the DC voltage drop across resistors in the collector circuits. Typical collector current of stage Q2 is 25-40 mA; Q3 and Q4, 30-60 mA. Transistor Q5 is tuned for maximum RF output into a 50-ohm dummy load.

Connect the negative (-) lead of your VOM or DVM TP3 and the positive (+) lead to test point TP1. Make sure the supply voltage is set to 13.2-volts DC. Set the slug in oscillator coil L1 all the way inside. Now back out the slug with a *plastic* align-

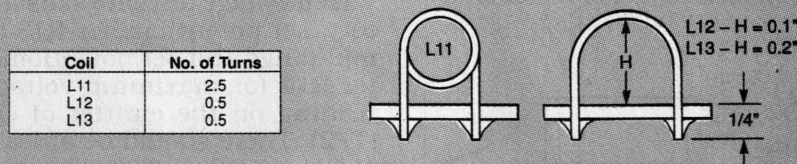
COIL WINDING DATA FOR THE MINI 900 TRANSMITTER



Wind coils L1-L5 using #22 enameled wire on an 8-32 screw. Form leads as shown in diagram. Scrape off enamel and tin wire ends as shown. Bottom of last winding is mounted 1/8-inch above PC board. Install coils on PC board using 8-32 screw as holder. Once soldered in place, remove screw and thread in 8-32 Cambion blue tuning slug.



Wind coils L6-L8 using #22 bare, tinned wire on an 8-32 screw. Form leads as shown in diagrams to allow 1/8-inch clearance for L7 and L8. Remove screw. Spread turns evenly to fit on PC board mounting holes and solder in place.



Wind coils L11-L13 using #22 bare, tinned wire on a 6-32 screw. Form leads as shown in diagrams. Remove screw. Spread turns of L11 evenly to fit on PC board mounting holes and solder in place. Solder coils L12 and L13 in place carefully observing height dimensions.



Toroidal coil L14 is wound on a 0.375-inch diameter ferrite core. Use fine sandpaper to remove sharp edges from the toroid-hole ends to avoid scraping of wire insulation when winding coil. Use #22 enameled wire (about 10-inches long). Pass the wire through the hole 13 times—each pass is a turn. Space turns evenly on toroidal form. Scrape off enamel and tin wire ends as shown. Install on PC board tight to surface.

coaxial cable is preferred. Cable type RG174 is also satisfactory, but tends to melt rather easily when soldered. Some sort of indicator on the load is necessary to see the RF output, but it does not need to be calibrated—a relative reading is satisfactory.

Restore power to the PC board. Readjust L2 and L3 as before. Now remove the negative meter probe from TP3 and connect it to TP5. The meter's positive probe is still connected to TP1 (13.2- volts DC).

Start with a range setting on the multimeter that will adequately show readings in the 1-volt DC range. Verify that the trimmers are preset. Adjust L4, L5, C12, and C13 for maximum meter reading. Then, readjust L2 and L3 to further increase this reading as much as possible. Repeat this procedure until no further improvement is noted. Capacitor C14 is not touched and is left as preset during this step. The reading obtained in this step may be typically 1 to 1.5 volts.

Repeak L4, L5, C12 and C13. Continue this process until no further increase in the meter reading can be obtained. The settings for the trimmer capacitors should not be too far from the original preset positions. If they are radically different, check to see if L4, L5, L6, and L7 are properly wound and that the correct parts are installed in the circuit board positions. Coil L4 has only 2.5 turns and may not securely hold the slug in position. Use a small drop of paraffin wax as a temporary cement to prevent the coil's slugs from falling out or shifting during handling and tune-up.

Adjust C14 for a dip (minimum reading at TP5) from the reading in the previous step. This dip in the reading may be around 30-50%. Do not worry about the exact amount as long as a significant dip is obtained. It should be fairly near the preset mesh, 10 to 30% or thereabouts. If the dip occurs at over 35% mesh, this indicates something has been improperly tuned or there is an error in assembly. Capacitor C14 should not be meshed more than 35%

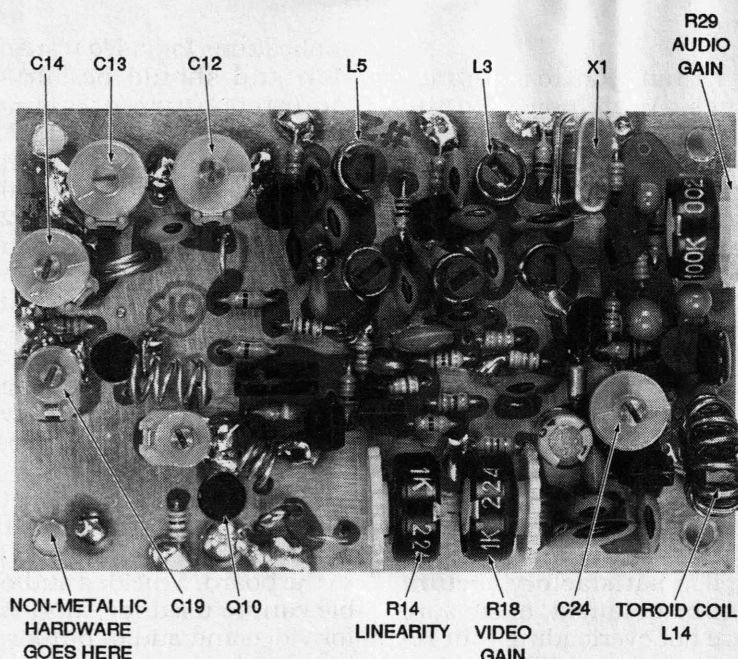
ment tool, while watching the meter. At some point the voltage reading on the meter should sharply jump up to a higher reading, then tune for a peak reading. Then back the slug of L1 out of the coil so the reading decreases about 5-10%. Next adjust the slugs of L2 and L3 for a maximum reading. Repeat adjustments of L2 and L3 until no further voltage increase is obtained on the meter. Next remove the power and set trimmer capacitors as follows:

C12 35% mesh
C13 20% mesh
C14 20% mesh

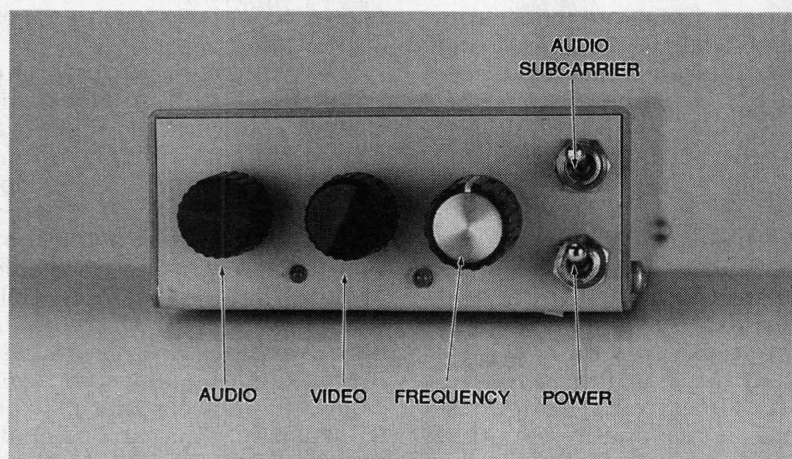
C19 35% mesh
C20 50% mesh

See Fig. 3 for locations of the above trimmer capacitors.

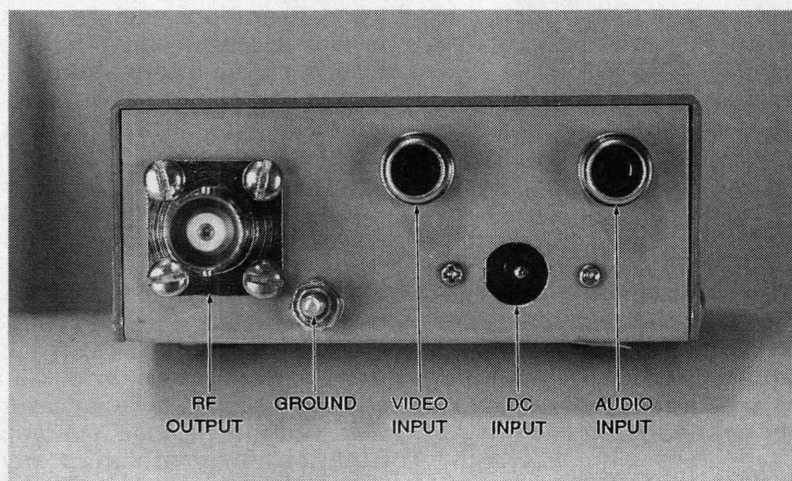
Connect a 50-ohm dummy load suitable for 900 MHz across the output of the RF power amplifier (junction C36 and L13, and foil ground.). Use 50-ohm coaxial cable and make the connection short. See Fig. 4 for proper cable-connection method. Connect the cable to the 50-ohm dummy load using a suitable connector to match your load (BNC or N-types, usually). Use proper adapters as required. Type RG188 or RG316



Component side of the PC board shown with all the parts installed and ready for cable connection to video and audio inputs and antenna. A few of the components are identified to facilitate compatible recognition with Fig. 3.



FRONT PANEL for the Mini 900 shows the operator's controls. The FREQUENCY control is a modification that was eliminated in the final prototype.



THE REAR PANEL is where all input and output connectors are located. A screw-terminal grounding post should not be ignored when setting up the transmitter.

as this indicates incorrect tuning.

Carefully and slowly adjust C19, while watching the RF output power meter and the power-supply current. A sharp rise in power-supply current and an indication on the RF power meter will be noted at some point around the preset position of C19, typically 30 to 40% meshed. *No other setting is acceptable.* The tuning elements are fixed in the PC board and therefore have close tolerances. Next, adjust C20 for maximum RF output. Then, go back to L4 and L5, and readjust them for maximum RF output, followed by re-peaking C12, C13, and C14 to further improve output. Reduce the supply voltage to 10-volts DC and verify there is still RF power output as seen at the indicator on the dummy load. About 20% or more of the previous output should still be obtained. If not, repeat previous steps starting with L2 until you achieve this.

You should be able to obtain RF output below 8.5 volts from the power supply before the RF power drops out completely and the RF power should vary smoothly with supply voltage, no sudden jumps or reversals at higher voltages, if the tune-up procedure was done correctly. There is some interaction between tuning adjustments. Also, the power-supply voltage can affect the value of transistor internal capacitances, resulting in some slight variation of optimum tuning settings over a voltage range. It is strongly recommended to perform initial tune-up at 13.2-volts DC, and make sure that all is functioning properly. If other supply voltages are to be used (9 to 15-volts DC) you should then re-tune the Mini 900 at the alternate supply voltage.

At 15 volts (maximum) power-supply voltage, the Mini 900 will have less tolerance to fault conditions such as over-voltage and high antenna VSWR (greater than 2:1). Conversely, at 9-volts DC (minimum) power-supply voltage, there is no room left to allow for lowered supply voltage (weak or exhausted batteries,

etc.). For this reason, 11 to 14-volts DC is suggested as an operating supply voltage.

About 0.5-watts RF output or more should be obtained with a 13.2-volts DC supply. At 10-volts DC, typically 0.2 watt; at 15-volts DC, typically 0.7-watt will be obtained. This is *peak envelope power* (PEP) and occurs on video signal's sync tips. Average power varies with video modulation and will be 20 to 60% of this figure, depending on the modulating signal. If, after all tuning adjustments are optimized, power output seems a little low, it is possible that the inductances of L12 and/or L13 are too high.

Check L12 and L13 by bringing a brass or copper screw near the coils, first L12 and then L13. The presence of the brass screw reduces the coils inductance. If the power output increases slightly, reduce the size of the corresponding inductor until this test fails to produce an increase. These inductors are electrically and physically small and there is a tendency for most builders to make them too large. Conversely, if no improvement is detected, leave them alone, and check C33, C34, C35, and C36 to see if they are correctly mounted and are not cracked or broken. Make sure that R14 and R18 are set for maximum voltage (at least 11.5 volts) at TP2.

The final adjustments

The last step is the checkout of the video modulator. You will need a suitable receiver or monitor capable of tuning to 902-928 MHz or to whatever crystal you have installed. Connect the transmitter to an antenna such that its radiating element is at least 5 feet from the PC board and at least 10 feet from the receiver. No antenna should be needed on the receiver. Connect a source of video (1-volt, peak-to-peak, 75-ohm, negative sync) to the video input of the transmitter (across the junction of R16- R17 and ground). Connect audio signal to audio input point (across R29). A VCR with a tape playing is an excellent video/audio

source.

Set R18 and R29 to mid-range settings. Apply power to the Mini 900 and adjust R14 so the RF power output drops to half its original level. Turn on the receiver and activate the VCR or other source of video and audio. Watching the picture, adjust R14 and R18 for a good stable picture, without rolling, or white or black level clipping. Adjust R29 for best sound. Set the receiver's volume at normal level during this adjustment. Readjust R14, R18 and R29 as needed.

Do not attempt any RF adjustments using the received picture. If a satisfactory picture cannot be obtained, make sure you are not overloading your receiver with excess RF signal or that you are not causing the RF signal to interfere with your source of video. Some video equipment may not work properly in strong RF fields so do not immediately assume the transmitter is to blame.

The video modulator in the Mini 900 is DC coupled. A drop in supply voltage due to battery depletion may cause sync clipping to occur. If this operating condition is expected, set the alignment adjustments at lowest expected supply voltage. An increase in supply voltage usually has no effect in this regard. The circuit design assumes a reasonably constant supply voltage, -1 to +2 volt variation at 12 to 14-volt levels. At 10 volts, no more than 0.75 volt drop would be advisable without readjustment of R14 and R18.

After alignment procedure is completed, it is a good idea to glue the tuning slugs in L1 through L5 in place, and to coat the toroidal coil L14. Use Duco cement or any clear lacquer for this purpose. Avoid pigmented paints as certain pigments can introduce losses into the coils. After the cement dries, a slight readjustment of the 4.5 MHz subcarrier frequency (C24) might be required.

Using the Mini 900

The Mini 900 ATV transmitter can be used in a variety of

applications for video transmission and should be suitably mounted. Current amateur band plans for 902-928 MHz suggest the use of two channels (picture carrier frequencies) for ATV use at 910.25 and at 922.25 MHz. This is not a legal requirement but a recommendation to follow that avoids possible interference with other amateur stations. Check the ARRL repeater directory or with local amateurs active in ATV to determine what picture carrier frequencies to use.

Also remember to maintain good RF practice and to shield all cables (video, audio and DC) to the board. Shielded audio cable can be used for short runs for video and audio, but for the antenna lead use a short length of RG188, RG316 or RG174 cable. These cables are very lossy so do not be surprised to experience a 20% power loss in a 2 or 3-foot cable.

Make sure to mount the transmitter in a shielded enclosure. If the overall weight is a consideration, a thin plastic enclosure wrapped with sheet copper foil might be used. Copper foil with a thickness of 0.001 or 0.0015-inch is usually available at hobby and craft stores. Solder all seams in the copper where possible. Keep the radiating antenna as far as you can from other devices that may be RF sensitive. This can be a challenge in R/C models. Use the best antennas possible. The Mini 900 may be limited to a low-gain antenna such as a simple quarter-wave whip. The 900-MHz receiver might require a higher gain antenna. Resist the temptation to use those cute little stub and rubber duckies when significant range is required. The larger types of antennas perform better, all things being equal.

The authors used a pair of corner reflectors for the range test and obtained a clean picture at over 5 miles, and a fair picture 12 miles away at favorable locations. Range is more a function of terrain than anything else. A corner reflector antenna consists of a half-wave

Continued on page 73

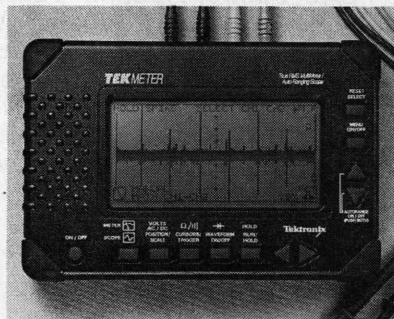
the line is used, and you do not have to stay on the same wire from one device to the next. The HS20 is also not polarity sensitive, so either lead of the semiconductor can be connected to either line in the loop. Again, you need to put one HS20 in the line for each device you want to protect including the modem. Any phone connected to the line without the HS20 in its line will be able to butt-in on your conversation, or on your data or fax transfer.

One HS20 bilateral silicon trigger switch is required for each phone and modem you intend to protect; the minimum quantity is two semiconductors for a home with one modem and one phone. If you are having trouble obtaining the HS20 bilateral silicon trigger switches from electronic parts suppliers, you can purchase them from SolarWorks, 2747 Wentworth Drive, Grand Prairie, TX 75052 for \$4 each, plus \$0.50 postage and handling per order. Ω

TROUBLESHOOTING

continued from page 38

each minute. As we listened to the machinery and watched the pulses, we heard a cooling fan switching just as each new spike occurred. Then we realized that the problem was the spike. It was fooling the PLC, and making it think that the temperature had dropped below threshold during one of the spikes, when the temperature dropped near 400 degrees.



THE TEKETER saved the day!

Now it's so obvious

A visual inspection of the thermocouple cables revealed that a portion of its shield had been worn away because of the vibration of a nearby cooling fan. Figure 6 shows some typical thermocouple hardware for those who are not already familiar with them. Replacing the cable got the annealing cycle back to normal, and the springs were springy again! We and our portable scope found and eliminated a costly factory problem.

The preceding examples illustrate just a few of the ways that a portable scope can be used to find and cure factory process problems. In addition to these simple cases, we can use the DVM and oscilloscope functions to find power-quality problems caused by dirty AC, bad power factors, and unbalanced phase power. We can also evaluate SCR- and triac-based motor controls for voltage, phase, and frequency conditions, factors governing efficiency. Finally, we can detect and cure sensor signal degradation with the simplicity of using a pocket DMM. Ω

ATV TRANSMITTER

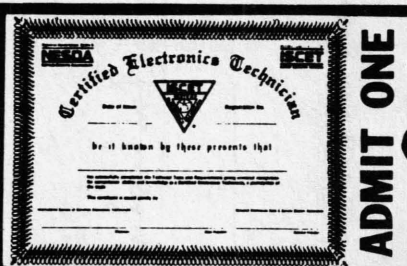
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dipole mounted near a reflecting surface consisting of two plane conducting surfaces at an angle, usually 60 to 90 degrees. The dipole is parallel to the line of intersection of these surfaces and 0.25 to 0.4 wavelengths away from the intersection. Typical gain of these antennas is 8 to 12 dB over a dipole.

A fair quality picture with minor snow requires better than 20 dB signal-to-noise ratio of the received signal. A 20 dB signal-to-noise ratio voice signal means very comfortable copy (assuming AM modulation). A fair picture means it has some snow, certainly useful for facial identification.

In the author's tests, the Mini 900 was mounted on the back of a corner reflector with 11-inch square sides at a 90-degree angle. The antenna was 30 feet above ground on a hilltop at the author's home, 12 air miles east of Glens Falls, NY. The receiver setup used a similar but slightly smaller antenna with a low-noise downconverter mounted on the back of the antenna. The matching downconverter for the Mini 900 ATV transmitter project and corner reflector antenna construction will be covered in future issues.

The downconverter was feeding a 12-volt DC TV receiver, the receiving setup located on the front seat of the author's vehicle. At best sites, the transmitter location was visible with high-power binoculars, so line-of-sight conditions were had, but a good picture was still obtained with some obstructions. However, in hilly terrain with heavy foliage, range was as short as half a mile when hills blocked the signal. R/C airplanes typically have line-of-sight conditions in most cases. Typical range with 3-inch whip antennas at the transmitter and receiver would be about one mile or so line-of-sight, less for obstructed paths. Use the best antennas possible for the situation you face. Ω



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WILLIAM SHEETS AND RUDOLF F. GRAF

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Our wireless link can transmit high-quality audio and video to any UHF TV channel. Low power requirements allow the transmitter to run on AA penlight batteries, and the PC board can be mounted easily in a small metal case, complete with batteries and a short antenna. The transmitter is crystal-controlled and it's easy to build and align. In addition, it uses low-cost, easy-to-get components, and it can be built for well under \$100.

Transmitter circuit overview

The block diagram of the wireless video link is shown in Fig. 1. The RF chain is fairly conventional. Its first stage is a crystal-controlled oscillator (Q1) with a frequency of 60 to 65 MHz, which is one-eighth the final output frequency. For example, our prototype used a crystal frequency of 60.40625 MHz, which gives a final output frequency of 483.25 MHz—the video-carrier frequency of UHF channel sixteen.

The oscillator produces a signal of about +6 dBm (4 milliwatts) that drives three stages of frequency doublers. The combined action of those doublers multiplies the input frequency by eight for a final output frequency of (nominally) 500 MHz. Double-tuned circuits are used between each stage to help reduce spurious outputs that might cause unwanted interference.

The video input signal (from your VCR, video camera, etc.) drives a video modulator (Q6 and Q7) that adds the video signal to the +12-volt line supplying power to the final



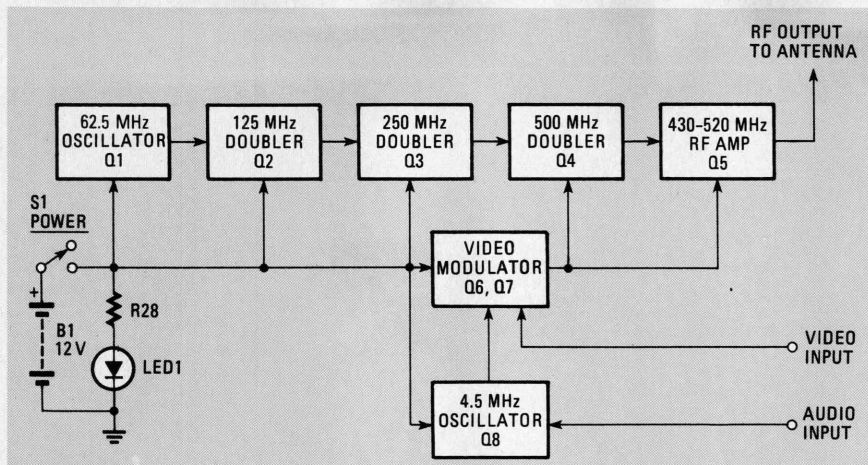


FIG. 1—BLOCK DIAGRAM OF THE TV TRANSMITTER reveals the overall simplicity of the circuit, which is composed of an oscillator, three frequency doublers, and video and audio modulators.

doubler (Q4) and the output amplifier (Q5). That method of modulation is similar to the way a conventional AM-radio transmitter is modulated. The video modulator has a nominal bandwidth of five MHz.

Audio input is applied to Q8, which operates as a VCO (Voltage Controlled Oscillator) running at a nominal frequency of 4.5 MHz to produce the modulated sound carrier. For simplicity, Q8 is a free-running oscillator, since the ± 25 -kHz frequency deviation that is required would be very difficult to produce at that frequency with a crystal-controlled oscillator. Besides, most TV sound systems will accept a ± 10 -kHz error in the sound-carrier frequency without producing undue distortion, and that greatly simplifies the circuitry required.

Calculating maximum range

The following equations may be used to calculate the maximum range you can expect from the wireless video-camera link. All logarithms in this equation (and in those following) are calculated in base 10; frequencies (f) are specified in MHz; and distances (D) are specified in miles.

The average TV receiver has a bandwidth (BW) of 4.0 MHz and a noise factor (NF) of 6 dB. For a snow-free picture, the carrier-to-noise ratio (C/N) should be 40 dB or better. Receiver sensitivity can then be calculated as the Minimum Desired Signal (MDS):

$$\text{MDS} = \text{NF} + 10 \log (\text{BW}) - 174 + \text{C/N}$$

Plugging values into that equation, we find that:

$$\text{MDS} = 6 + 10 \log (4 \times 10^6) - 174 + 40$$

Therefore, $\text{MDS} = 62 \text{ dBm}$, or $0.794 \times 10^{-3} \times 273 = 216 \text{ millivolts}$. So for a transmitter power of $+15 \text{ dBm}$ and 2 isotropic antennas, that allows a Path Loss (PL) of 87 dB.

$$\text{PL} = 37 + 20 \log (D) + 20 \log (f)$$

Isolating the distance factor we see that:

$$20 \log (D) = \text{PL} - 37 - 20 \log (f)$$

$$20 \log (D) = 87 - 37 - (20 \times 2.7)$$

$$\log (D) = -4/20 = -0.2$$

So the maximum range obtainable should be:

$$D = 10^{-0.2}$$

$$D = 0.6 \text{ mile} = 3100 \text{ feet}$$

However, you wouldn't really be able to get a range of 3100 feet because of reflections, loss from "dead spots," terrain loss, and obstacle shielding. However, a distance of several hundred feet is easily possible using a $+12 \text{ V}$ supply.

Detailed circuit description

The complete schematic of the wireless video link is shown in Fig. 2; we'll discuss each stage in detail. Transistor Q1 is a common-base (or Colpitts) oscillator biased by resistors R1, R2, and R3. Inductor L4 and capacitors C3, C4, C5, and C8 form a circuit that is tuned to the frequency of the crystal.

The crystal is series-resonant at some frequency between 60 and 65 MHz, so it appears as a low impedance (50 ohms or less) at that frequency. Therefore Q1 will have sufficient gain as a common-base amplifier only at the resonant frequency of the crystal. Hence the signal developed at the junction of C4 and C5 will be amplified by Q1 only if that signal is at the same frequency as the crystal. At that frequency, Q1, has sufficient gain to oscillate because the ratio of the voltage initially developed between C4 and C5 to that at Q1's collector is greater than unity.

Capacitors C3 and C8 complete the tuned circuit; they also form a voltage divider that feeds the base of Q2 about one volt of the signal from Q1. Transistor Q2 functions as an overdriven amplifier that distorts its input signal and thereby produces harmonics of the input frequency. The second harmonic (120 MHz) is the

PARTS LIST

All resistors $\frac{1}{4}$ -watt, 5% unless otherwise noted.

R1, R5, R11—22,000 ohms
R2, R6, R12, R16—4,700 ohms
R3, R7, R13—220 ohms
R4, R9, R10, R18, R19, R21, R25—100 ohms
R8, R14, R33—10 ohms
R15, R26, R35—100,000 ohms
R17, R28, R30—2200 ohms
R20, R22—470 ohms
R23, R27, R34—10,000 ohms
R24—3300 ohms
R29—82 ohms
R31—1000 ohms, linear taper potentiometer
R32—10,000 ohms, audio taper potentiometer

Capacitors

C1, C6, C15, C32, C38—0.01 μF 50 volts, cermaic disc
C2, C7, C9, C14, C16, C23, C26, C29, C30—470 pF cermaic disc
C3—33 pF, 5%, mica
C4, C19—15 pF, 5%, mica
C5—56 pF, 5%, mica
C8—82 pF, 5%, mica
C10—18 pF, 5%, mica
C11—2 pF, $\pm 1 \text{ pF}$, mica
C12—24 pF, 5%, mica
C13—39 pF, 5%, mica
C17—8 pF, 5%, mica
C18, C24—1 pF, $\pm 1 \text{ pF}$, mica
C20—12 pF, 5%, mica
C21, C27—47 pF, 5%, mica
C22, C25, C28—1–8 pF, polystyrene trimmer
C31, C33, C39—8.2 μF , 20 volts, tantalum
C34—470 pF, 5%, mica
C35—220 pF, 5%, mica
C36—5–60 pF, polystyrene trimmer
C37—100 pF, 5%, mica

Semiconductors

Q1–Q3, Q6—2N3563
Q4—2N3564
Q5, Q7—2N3866
Q8—MPF102
Q9—2N3565
LED1—standard red LED
D1—1N757 9V Zener
D2—MV2117 varactor
D3—1N4002

Other components

J1—Video camera jack
J2—BNC jack
J3—Coaxial power jack
L1—6.2 μH (see text)
L2, L3—0.074 μH (see text)
L4—0.15 μH (see text)
L5, L6—0.035 μH (see text)
L7, L8, L9—0.018 μH (see text)
L10, L11—5.6 μH
S1—SPST miniature toggle
XTAL—Crystal (see text)

Miscellaneous: Aluminum project case
AA-penlight cell battery holders

Note: The following are available from North Country Radio, P. O. Box 53, Wykagil Station, New York 10804: Etched and drilled PC board, ferrite cores for coils L1 and L4, and crystal for Channel 14 or 15, \$32.50; Complete set of all parts that mount on PC board, \$69.95. In either case, be sure to specify channel.

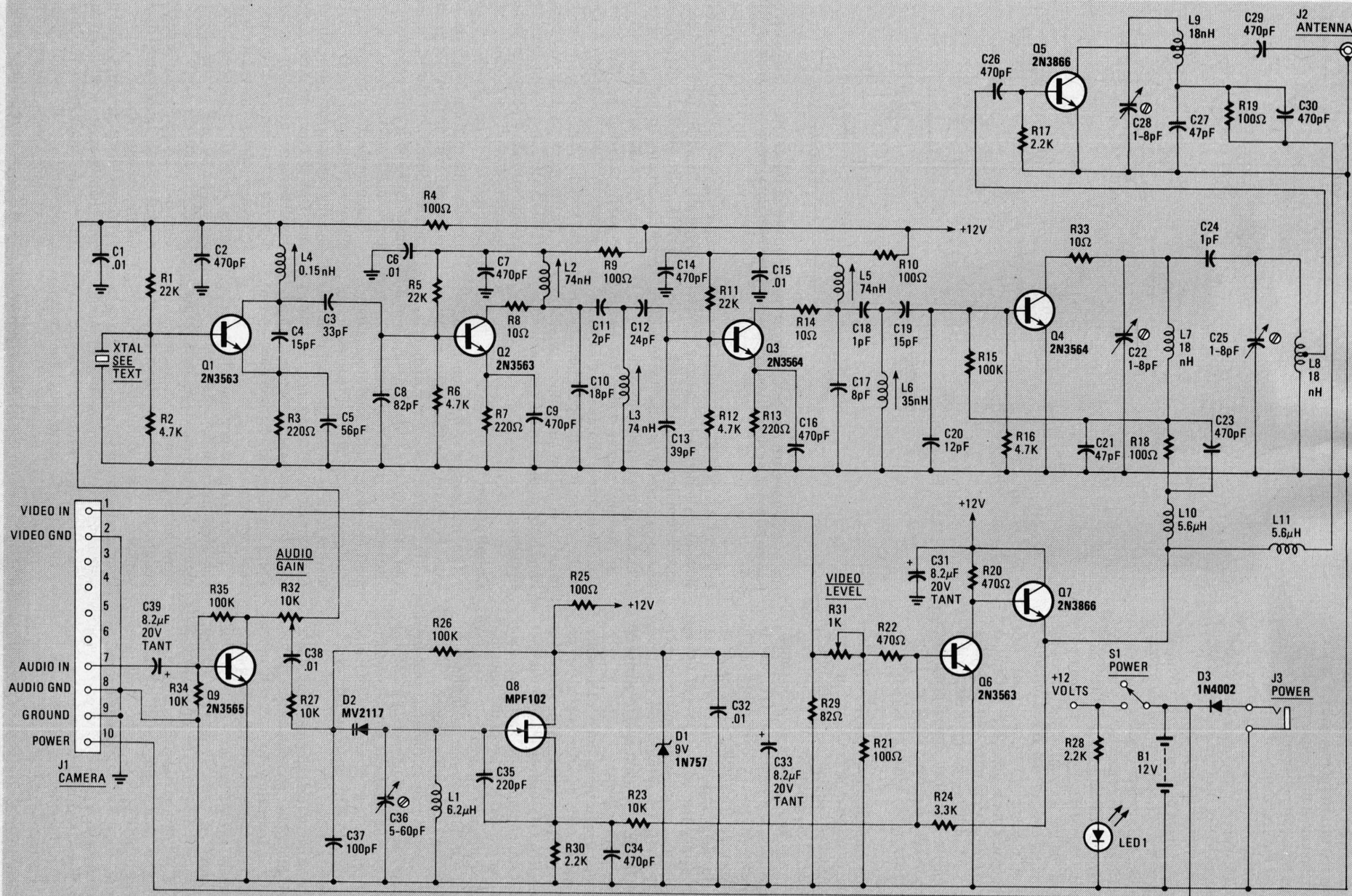


FIG. 2—COMPLETE SCHEMATIC OF THE TV TRANSMITTER is shown here. The circuit is straightforward, but see the text for information on winding coils L1–L9.

frequency we're interested in; L2 and C10 are tuned to that harmonic, and C8 is also series-resonant at that frequency. The additional base current supplied by C8 also improves Q2's efficiency of oscillation. A double-tuned circuit is provided by C11, C12, C13, and L3; those components filter harmonics higher than the second, as well as the 60-MHz fundamental.

Another stage of frequency doubling is provided by Q3, which operates very much like Q2, except that the tuned circuits at its input resonate at approximately 125 MHz, and its output circuits resonate at about 250 MHz. Again, Q4 operates like Q3, taking account of the values of the components in the tuned circuits. However, note that no emitter-bypass capacitor or resistor is used. It is difficult to get good bypassing in the 430–500-MHz range with ordinary components, and it takes only a very small impedance in the emitter to kill the power gain of that stage. Therefore, the emitter is directly grounded.

Power amplification is provided by Q5; it receives its drive from Q4 and the double-tuned VHF circuits composed of L7, L8, C22, C24, and C25.

Both Q4 and Q5 receive their supply voltage from the emitter of Q7, which supplies +4.5 volts with no input signal to Q4 and Q5. That voltage has positive-synch-tip video superimposed on it. The video gain provided by Q6 and Q7 is about eight, and the bandwidth is greater than 10 MHz. Those transistors are capable of driving a 75-ohm load to a level of 10 volts peak-to-peak.

Negative-synch input video from a camera, VCR, etc., is DC-coupled to the junction of R21 and R22. Video bypass is provided by C31. Gain and Q-point are set by R24; potentiometer R31 acts as a video gain control; and R29 keeps the input impedance around 75 ohms.

FET Q8 functions as a Hartley-type VCO with a free-running frequency of 4.5 MHz; C36 is used to fine-tune that frequency. Feedback is provided by C35 and C34; D2 is a varactor (variable-capacitance) diode. It is biased at about nine volts by R25, R26, and Zener diode D1, which also biases Q8.

The varactor (D2) changes capacitance at the audio rate, and that causes the oscillator's frequency to vary. In other words, the varactor provides frequency modulation (FM). Audio is fed to D2 via R27 and C38, and C37 provides RF bypassing at 4.5 MHz. The small value of C38 (0.01 μ F) provides pre-emphasis to the audio. Audio pre-amplification is provided by Q9; R32 is the audio gain control. The 4.5-MHz FM signal from Q8 is summed with the video signal through R23. The sound-level carrier may be varied by changing R23 as necessary.

The power supply is a 12-volt battery pack composed of eight AA alkaline

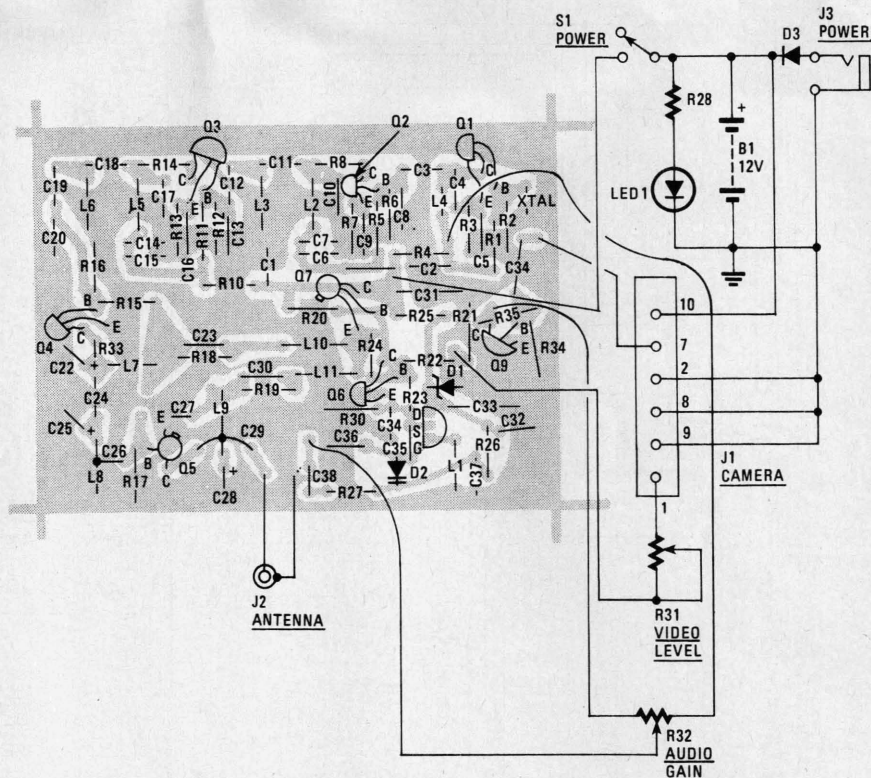


FIG. 3—COMPONENTS ARE MOUNTED ON THE PC BOARD as shown here. Note how R31 dangles from the PC board. We suggest you use a more secure method of mounting that potentiometer than we did.

cells; 10 AA Ni-Cd cells could be used instead. Alternatively, external power may be coupled in through jack J3. The power-on condition is indicated by LED1, which is current-limited by R28.

If, for any reason, video bandwidth must be reduced, simply add a small capacitor across R24. It should have an impedance equal in magnitude to R24's impedance at the highest intended video frequency.

Construction

Our transmitter is built on a single-sided PC board; the foil pattern is shown in "PC Service" elsewhere in this magazine. You may use either 0.8- or 1.6-mm G-10 glass-epoxy PC-board material. For most low-power, solid-state applications, G-10 epoxy is suitable for frequencies at least as high as 1000 MHz. That material has high electrical resistance, low electrical loss, mechanical rigidity, dimensional stability and low water-absorption characteristics.

The majority of construction projects published the past few years have been digital—low frequency—projects. Since many hobbyists have never attempted to build a high-frequency project like our transmitter, a few words about UHF construction techniques are in order. So bear with us for a few moments if what we say below is old hat to you.

First, be careful to select the proper components. For example, at low frequencies (under about 10 MHz), most ce-

ramic bypass capacitors work fairly well. Typical 0.01 and 0.1 μ F bypass capacitors in TTL and CMOS circuits fail to do their job as operating frequency is increased to about 25 or 30 MHz. That may be corrected by reducing the value of those capacitors to 0.001 μ F, and by keeping the lead lengths short—zero, if possible.

As the operating frequency approaches 100–150 MHz, bypassing again becomes a problem. At that point even smaller values (47 to 470 pF) are required, and leads must be very short. At even higher frequencies, up to about 500 MHz, chip capacitors, which have no leads, are soldered directly to the PC board. The moral is: Keep leads very short—do not use 1/4-inch leads on bypass capacitors.

Common electrolytic capacitors generally work well up to 10 MHz, depending on value and physical size. Mica capacitors (DM-15 type) are pretty good up to the VHF range, depending on value, but they can have from 5 to 15 nanohenries of series inductance. That inductance can increase apparent capacitance. For example, a capacitor with a value of 47 pF at 30 MHz may measure 82 pF at 150 MHz. That must be taken into account when designing tuned circuits at the frequencies we are dealing with.

Typical 1/4-watt resistors of 15 to 1000 ohms behave pretty well up to about 250 MHz or so. Values below 15 ohms tend to appear inductive, and values above 1000 ohms tend to appear capacitive. That can cause a shift in impedance, and, there-

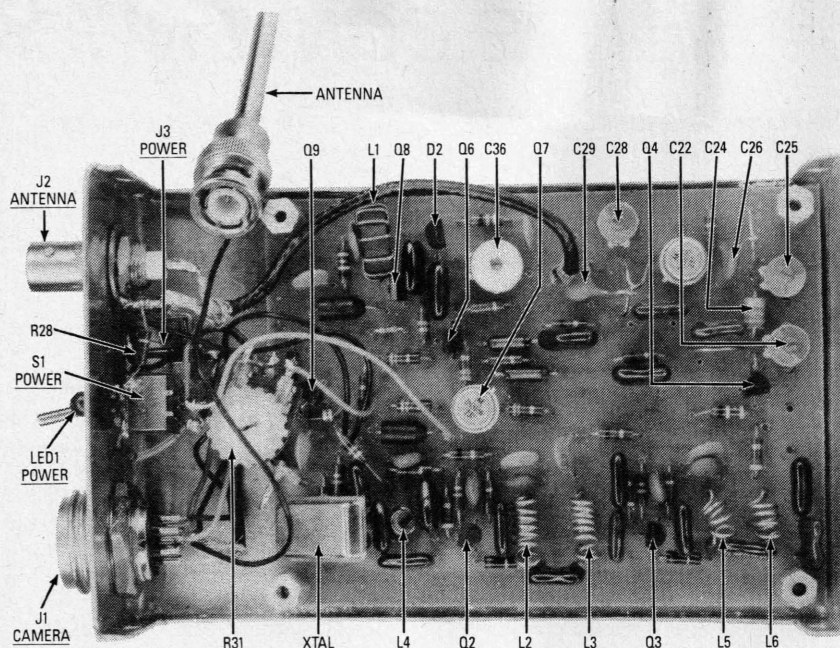


FIG. 4—MOST COMPONENTS ARE STANDARD, but note how C26 and C29 connect to L8 and L9, respectively. Also note the unequal turn-spacing of L2, L3, L5, and L6, which was brought on by tuning those coils. Finally, note the small slug L4 is wound around, and L1's toroidal core.

fore, the resonant frequency, of a tuned circuit. Stray inductive and capacitive reactances cause that shift.

Also, remember that every PC trace has inductance and capacitance. A PC pad may have, depending on size, several pF of capacitance to ground or to another PC pad. A long PC rail may act as an RF transmission line or even as a resonant circuit. It may also appear as an additional turn on a coupling transformer, by which unwanted signals might be radiated to other components or PC traces.

We hope you're not scared by all those cautions. We just want you to be aware of some of the problems involved in RF design and construction. But if you take care to duplicate our prototype exactly, the chances are that you'll have no problem getting your transmitter working. There is no "black magic" involved—only common sense and careful construction. Just be sure to use the same or equivalent parts—be careful!

After you have your components together, check the PC board over for shorts and opens, and make sure that the copper is clean and shiny. Then, referring to the parts-placement diagram in Fig. 3, insert and solder the components, starting with those that have the lowest profile (the resistors and diodes), and working up to the electrolytic and trimmer capacitors. Remember to cut all leads—especially the capacitors' leads—as short as possible. Don't overheat the semiconductors when soldering them to the PC board.

Now you're ready to wind and install the RF coils. Spread the turns of each coil evenly, but don't worry about spacing those turns perfectly, since the coils will

be compressed and expanded later when you tune up your transmitter. You can see in the photo (Fig. 4) how our prototype's coils turned out.

- L1 is eight turns of 22-gauge wire wound on a Ferroxcube 768T188 toroid core made from 4C4 ferrite.

- L2 and L3 are seven turns of 22-gauge wire wound on a #26 drill bit. Of course, remove the drill bit after the coil is completed.

- L4 is wound around a standard 10-32 screw thread. The screw should be removed after the coil is completed. Then L4 should be fitted with a ferrite slug. You may be able to find an appropriate slug in an old TV, radio, or CB radio. Best results will be obtained when the slug is taken from a circuit that operates in the 25–100-MHz range, such as a TV IF circuit, or the front end of an FM radio. But it's not really critical, and almost anything should work (provided it's made of ferrite) as a last resort. If necessary, L4's diameter, and number of turns, could be changed to fit the slug you have.

- L5 and L6 are three turns of 22-gauge wire wound on a #26 drill bit.

- L7, L8, and L9 are merely 1.5 cm loops of wire wound on a 3/8-inch form and sol-

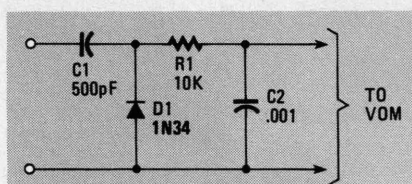


FIG. 5—RF-PROBE may be used with any VOM or DVM that has sensitivity greater than 20K ohms/volt.

dered to the PC board. One end of capacitor C26 is mounted in the normal fashion, and the other end hangs from the approximate mid-point of L8's loop. Similarly, C29 is mounted from the board to the mid-point of L9; the lead then continues to the pad near the collector of Q5.

- L10 and L11 are standard 5.6 μ H, 10%-tolerance chokes obtainable from the J.W. Miller Corporation, etc. They could also be wound from 36-gauge wire on 1/2-watt, 1-megohm carbon resistors if desired.

Last, install the transistors, making sure that they are oriented correctly and that their lead length is minimized.

We use a 10-pin camera jack for J1, but feel free to substitute whatever connectors you need. If no sound carrier is needed, R23, R25, CR2 and all other components associated with Q8 and Q9 can be omitted. Doing that will not affect the operation of the video portion of the transmitter. They can be added at a later date should audio transmission become necessary.

We chose not to leave space on the PC board for the audio- (R32) and video-input (R29, R31) components, since those components were unnecessary in our application. We used fixed-value resistors for R31 and R32, but small potentiometers could be mounted to the case and wires run to the PC board. If no gain control is necessary, R32 should be replaced by a fixed 10K resistor, and C38 should be connected directly to the collector of Q8.

Solder the coils to the PC board now, and solder short interconnecting wires from the board to the chassis components. Before applying power, check over your work: Make sure no solder bridges exist and make sure that all polarized components are correctly oriented.

Testing and alignment

The following equipment is necessary to align the transmitter:

- VOM or DVM having sensitivity of at least 20,000 ohms/volt
- RF probe
- Video source (VCR or camera)
- TV set or monitor
- 50-ohm dummy load

If an RF probe is not available, you can use the circuit shown in Fig. 5 with your voltmeter. If you have no 50-ohm dummy load, you can use a 51-ohm, 1/4-watt resistor. One handy gadget to have is a tuning wand: a plastic rod with a ferrite slug at one end and a brass slug at the other. The inductance of a coil can be increased or decreased by placing the ferrite or the brass slug, respectively, near the coil.

Apply power and check for +12V at R4, R9, R10 and the collector of Q7. Check for +9V at the drain of Q8. Then check for 1.0 to 1.5 volts at the emitters of Q1, Q2, and Q3. Check for +4.5 to

continued on page 114

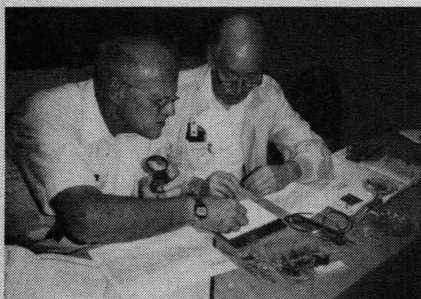
HIGHLIGHTS OF THE NPE CONVENTION

The unification of NESDA and NATESA was just one of the highlights of the 1985 National Professional Electronics Convention.

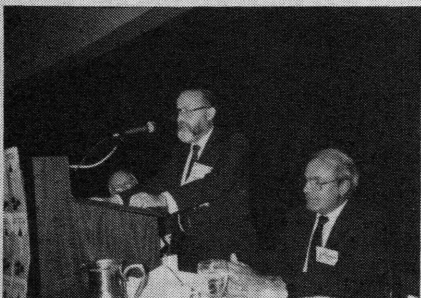
WHILE EACH YEAR'S NATIONAL PROFESSIONAL electronics convention (NPEC) is a worthwhile and memorable experience, 1985's session, held in August in Hartford CT, was particularly significant for the electronics industry. At that meeting a consolidation agreement between the National Electronics Sales and Service Dealers Association (NESDA) and the National Association of Television and Electronic Servicers of America (NATESA) was unanimously approved by NESDA. Later in the month, NATESA, at its convention held in St. Charles, IL, also unanimously approved it.

Joining together

Under the agreement, the Chicago-based NATESA organization will become the NESDA state affiliate in Illinois as of January 1, 1986. At the same time, NESDA will accept and perpetuate the heritage of NATESA as its own. For two years, current NATESA members who reside outside of Illinois may choose to either remain members of the Illinois association or to join the NESDA affiliate in their home state. Current NESDA members residing in Illinois have two years to join the new Illinois affiliate.



GETTING INVOLVED. Murray Barlowe, CET, and Ed Kimmel, CET, participate in the Electronics Instructors Conference, one of the many seminars at the Hartford NPEC.



MAN OF THE YEAR. Larry Steckler, publisher of *Radio-Electronics*, Chairman of the Electronics Industry Hall of Fame, and NESDA's Man of the Year speaks at the Hall of Fame Banquet. Roger Companion, NATESA Vice President, is seated at the right.

Former officers of NATESA will be recognized equally with those of NESDA. The founder of NATESA, the late Frank J. Moch, who, at the convention, was inducted into the Electronics Industry Hall of Fame, will be perpetually honored as the father of both organizations.

The current presidents of both organizations highly praised the new agreement. Said continuing NATESA president Tom Leeney, "We're proud to contribute our part to a vibrant NESDA organization, and this agreement shows that the people in our industry have the ability to work together for the good of all." Dorothy Cicchetti, the newly-elected president of NESDA, added, "With the addition of the diligent and loyal NATESA members, NESDA will become an even more dominant force for good in our profession."

Elections and awards

A spirited contest led to the election of Dorothy Cicchetti, of Flushing, NY, as NESDA president.

The other NESDA officers elected included: Robert Harrell, Johnson City, TN, 1st National Vice President; Clifford A. Shaw, Columbia, VA, Secretary; and Dick Scott, CET, Olympia, WA, Treasurer. Regional Vice Presidents are Art Van Sicklin, Hartford, CT; Faust Guarino, CET, Oceanside, NY; Jim Teeters, CET, Norfolk, VA; Gene Dillingham, CET, Louisville, KY; Bob Mesa, CSM, Parma, OH; Floyd Hack, CET, Taft, TX; Gennie Randel, Greensburg, KS; Vince Hostetler, CSM, Grand Junction, CO; Ken Duncan, CSM, Antioch, CA; and Bob Villont, CET/CSM, Tacoma, WA.

The new ISCET (International Society of Certified Electronics Technicians) officers are: Jim Parks, CET, Orlando, FL, Chairman; Don Winchel, CET, Smart-

ville, CA, Vice Chairman; Hal Robbins, CET, Van Nuys, CA, Secretary; and Earl Tickler, CET, Baltimore, MD, Treasurer. Bob Villont will continue as the NESDA representative to the ISCET board of governors.

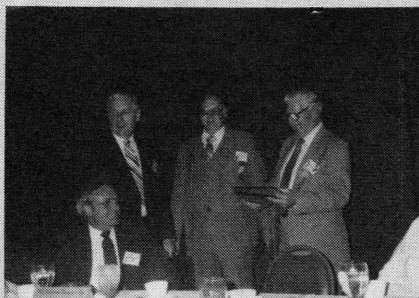
Turning to the awards, as previously mentioned, the late Frank J. Moch was inducted into the Electronics Industry Hall of Fame. In addition, Larry Steckler CET/EHF, Publisher, Gernsback Publications (including, of course, **Radio-Electronics**) and Chairman of the Electronics Industry Hall of Fame was named NESDA's Man of the Year; he was also presented with ISCET's Continuous Service Award.

Other major awards included: Ken Duncan, CSM, Officer of the Year; Cliff Shaw and Chic Young, P.E./CSM (tie), Outstanding Committee Chairman; Don Erwin, CSA, Outstanding State President; Ed Erich, CSM, Outstanding Local President; Martin Fleming, CET, Technician of the Year; and Lester Dodd, CET, 1985 Friend of ISCET.

Rounding out the event was an outstanding slate of professional seminars, a trade show, and an unusually rich assortment of memorable social functions. **R-E**



CELEBRATING THE SIGNING of the momentous agreement consolidating NESDA and NATESA are NESDA President Dorothy Cicchetti (left) and NATESA President Tom Leeney.



OUTGOING NESDA PRESIDENT George Bluze (standing left) presents certificates recognizing the industry contributions of Paul Kelley (standing center), Robert Haskins, and Robert Read. Tom Plant CET (standing right) accepted the certificate for Haskins and Read.



JEROME MOCH, (standing left) son of the late Frank J. Moch, accepts the plaque designating the induction of his father into the Electronics Industry Hall of Fame. Making the presentation is Wallace Harrison, Director of Communications of NESDA/ISCET.

VIDEO LINK

continued from page 55

+5.5 volts at the emitter of Q7, and check for +4.5V at the collectors of Q4 and Q5. The emitter of Q7 may show 4-7 volts—that's OK. Also, Q7 will normally run somewhat warm to the touch.

If all DC voltages check out, you can begin alignment. First, check the voltage at the collector of Q5. As you ground the base of Q6, that voltage should increase to over 10 if the modulator is working.

Next, connect the RF probe to the junction of C3 and C8. Tune L4 until Q1 starts to oscillate, as indicated by a reading of 0.5 to 1.0 volts on the VOM. Now connect the RF probe to the base of Q3 and spread or squeeze the turns of both L2 and L3 to obtain maximum voltage. Now connect the RF probe to the base of Q4 and adjust the turn spacing of L5 and L6 for a maximum voltage. Now, re-adjust L2, L3, and L4 to maximize the voltage reading. Connect the voltmeter to the base of Q5 and adjust trimmers C22 and C25 for maximum reading. Re-adjust L5 and L6.

Now connect the dummy load to J2, and connect the RF probe across that load. Adjust C22, C25, C28 and the tap positions on both L8 and L9 until no further increase in output can be observed.

If one stage has no output, the cause may be misalignment of (or circuit trouble in) the preceding stage. It may be necessary in some instances to add or subtract a turn from L2, L3, L5, or L6 if proper adjustment cannot be obtained by adjusting turn spacing. If a coil's turns are spread far apart, and the alignment calls for even more spacing, a turn should be removed. On the other hand, if a coil has to be "crushed" together, add a turn. (But make a new coil rather than splicing on more wire.)

Now tune the TV receiver to the transmitter's output frequency. You should see a blank raster when the transmitter is turned on. Connect a video signal to the input jack (J1) and adjust video level so that a stable picture is obtained. Increase the video level until tearing or rolling develops; that indicates sync-tip compression. Retune the Q4 and Q5 stages for best picture quality without tearing.

Adjust C36 for clearest audio while applying an audio signal to C38. Be careful not to overdeviate or misadjust the 4.5-MHz oscillator frequency. A frequency counter can be helpful for that. Connect it to the emitter of Q7 and adjust C36 for a frequency of 4.500 MHz with no video input (short R21 to ground).

That completes alignment and testing of the wireless video link. A very short antenna (about one inch) connected to a 51 ohm resistor makes a good limited-range antenna. For greater range, a six-inch whip antenna could be used. **R-E**

CURING EMI

continued from page 74

That isolating impedance is either an inductor (with a value between 50- and 500 μ H) in power-line hybrids, or, as shown in Fig. 9, a resistor (whose value is typically a few tens of ohms) in signal-line applications.

Designing transient suppressors

The most severe source of power-line transients is a lightning stroke. Therefore, most suppressors are primarily designed to protect against lightning. Most of the energy associated with a lightning stroke is in the form of common-mode current and voltage. However, due to unequal wiring parameters and the fact that common-mode suppressors are not perfectly matched (do not suppress equally), some differential-mode energy will also be present. Thus, a lightning-protection network should incorporate both common- and differential-mode protection.

Varistors, which are commonly used as suppressors on household AC power lines (117 volts), are available from a number of electronics suppliers. When selecting a varistor for a particular application, you should consider the nominal operating voltages and currents that the circuit will see, and the maximum surge voltage and current that are likely to be encountered. In addition, the suppressor must be rated to withstand the energy that it will need to dissipate. In the case of 117-volts AC, as stated previously, the most powerful transient that is likely to occur will have a peak voltage of 6000 and a peak current of up to 200 amperes; that's a fairly sizeable wallop. If the suppressor clamped that surge voltage at 500, the amount of energy that the suppressor would need to absorb would be approximately 0.8 joule. Lower clamping voltages will mean that the suppressor will have to absorb more energy.

In addition to power lines, varistors designed to protect low-voltage circuits (as low as 5 volts) are available. The selection of such a varistor follows the same general procedure as above.

Silicon avalanche suppressors also are commonly used to protect low-voltage systems. For instance, to protect sensitive CMOS devices on a PC board, a 5-volt SAS could be connected between the power and ground connections on the board. Data lines similarly could be protected by installing SAS's between the lines and ground.

The suitability of an SAS for a particular application is determined by its power rating. Those power ratings range from 500 watts to several kilowatts. An SAS with a power rating of 1000 watts, that is designed to clamp the voltage at 5, can handle surge currents of up to 200 amperes. **R-E**

SATELLITE-TV RECEIVER

continued from page 60

At this point you are ready to attach a TV (or a video monitor) and an audio amplifier to the receiver. CAUTION: Disconnect the receiver from the AC voltage source while hooking up everything. After everything is hooked up, reconnect the receiver to the AC voltage and follow these steps:

1—Turn on the receiver and adjust the TRANSPONDER TUNING control to receive a picture. Adjust the SKEW control and the POLARITY switch for the best picture.

2—Adjust C71 for maximum contrast in the picture. Gently compress and expand L2's coils slightly while observing the picture on several different channels. Adjust L2 for the best picture.

3—Adjust R109 for the best picture. Video level is controlled by that potentiometer, as is contrast. If it is adjusted to too high a value, a buzzing sound may be heard in the audio when lettering appears on the screen.

4—Set the SUBCARRIER TUNING control to the center position, and set the BANDWIDTH switch to WIDE. If no sound is heard, adjust R112 slightly. If nothing but noise is heard, adjust C72 until the audio comes through. That can be a "touchy" adjustment. Get it close, and then try fine-tuning the front-panel control. Once the sound is heard, readjust R112 for best audio.

5—Aim your dish at Satcom F-3, and tune in the appropriate transponder for either WTBS or WGN. Set the BANDWIDTH switch to NARROW, and then slowly turn the SUBCARRIER control counterclockwise from center. Several FM-radio programs should be heard. Adjust R111 for the best sound.

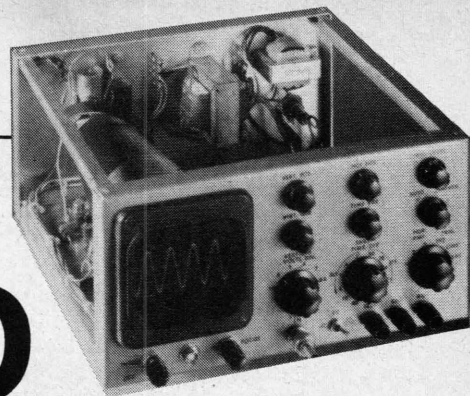
6—Adjust R107 for full-scale meter deflection when receiving the strongest station in your area.

7—Trimmers R102 and R104 may need to be adjusted slightly in order to make R103 correspond with the markings on the front panel. Set the TRANSPONDER TUNING control to the number of the lowest transponder channel received in your area and adjust R104 for best reception. Then set the panel control to the number of the highest channel in your area and adjust R102. Those adjustments will interact slightly, so go back and forth until both channels come in correctly.

At this point you can sit down and relax: you've got a fully-functional satellite-TV receiver! We have stuck pretty much to the basics in this article, but if you need more information, be sure to consult the manuals that accompany your LNA, down-converter, feedhorn and dish. Also, see "Installing Your TVRO," which appeared in the June and July 1985 issues of **Radio-Electronics**. **R-E**

BUILD THIS

TRIGGERED OSCILLOSCOPE



This easy-to-follow design lets you keep cost low by using a CRT of your choice. Its operational feature is a continuous zero baseline.

DANIEL METZGER and DENNIS PERRY

ONCE A TECHNICIAN HAS EXPERIENCED troubleshooting with a calibrated DC lab scope, he'll probably want to keep that scope probe close at hand most of the time he's at the service bench. Transistor base-emitter voltages, collector saturation voltages, and IC logic levels can be checked as easily as power-supply lines while the operating signals are present. No other instrument provides that simultaneous readout of bias and signal conditions.

Two factors have conspired to keep that scope probe out of the hands of most experimenters. The first is cost, which approaches \$200—even for a kit. That problem is easily solved by simplified design. The scope described here can be built from standard parts for \$100, and considerably less if the junk box is well stocked. Yet it boasts a 2-MHz bandwidth and 10 mV-per-division vertical sensitivity.

The second factor is the annoyingly frequent need to lay down the probe, reach over to the scope, throw the input switch from DC to GROUND, check the position of the zero-volt baseline, and throw the switch back to DC. That problem is handled by incorporating a circuit that provides a continuous display of the DC ground level at a brightness level lower than that of the signal display.

How it works

The operation of the scope as a whole is best understood from the block diagram, Fig. 1. The vertical attenuator and amplifier provide a replica of the input signal, both AC and DC, at the approximately 100-volt level needed at the deflection plates of the CRT. The electronic baseline switch interrupts the signal and grounds the amplifier input for ap-

proximately 3 ms each 15 ms, thus providing a 1/5 duty-cycle baseline display at a rate of about 60 Hz—too fast for the eye to perceive the flicker.

A separate trigger amplifier is fed from a point ahead of the electronic baseline switch to preserve continuity of sweep

triggering. A Schmitt trigger produces squarewaves in sync with the input signal, and a differentiator produces sharp spikes from the edges of the squarewaves. The negative spikes initiate a linear ramp that always starts at the same selected point on the input AC wave-

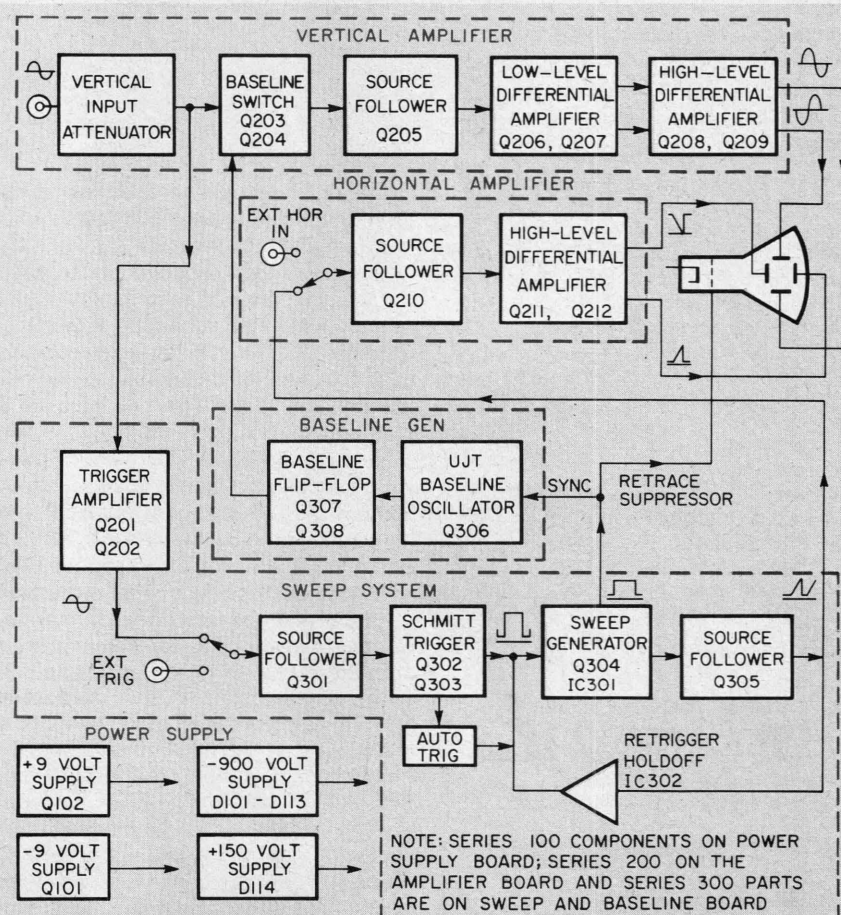


FIG. 1—BLOCK DIAGRAM of the zero-baseline scope. Operation is somewhat similar to a dual-trace scope with the baseline considered as the second trace.

form. That ramp is applied to the horizontal amplifier to produce the calibrated time sweep. A source-follower provides a low output impedance for the ramp, and an op-amp comparator holds off further triggering signals until the ramp voltage returns to zero.

An auto-trigger circuit senses when the Schmitt trigger is not switching and immediately applies a voltage to the ramp generator commanding continuous ramps, thus providing sweeps for the display of DC voltages.

A UJT baseline oscillator running at approximately 60 Hz is synchronized to the sweep generator to insure that the switching from signal to baseline will always occur during a retrace of the sweep. The baseline flip-flop drives the baseline-switching FET's at the input of the vertical amplifier.

The CRT cathode is operated at -900 volts to accelerate the electron beam toward the CRT face. Deflection sensitivity and hence calibration depend upon that voltage, so it is regulated by a string of 180-volt Zener diodes. Vertical and horizontal position and sweep time depend upon the 9-volt supplies, so they are transistor-regulated. The +150-volt supply serves only differential amplifiers, and their inherent common-mode rejection makes regulation of that supply unnecessary. We shall now proceed to a detailed description of each functional block.

Vertical attenuator: Voltage dividers R_A and R_B (Fig. 2) reduce the input signal to a maximum of 0.32 volt (8 divi-

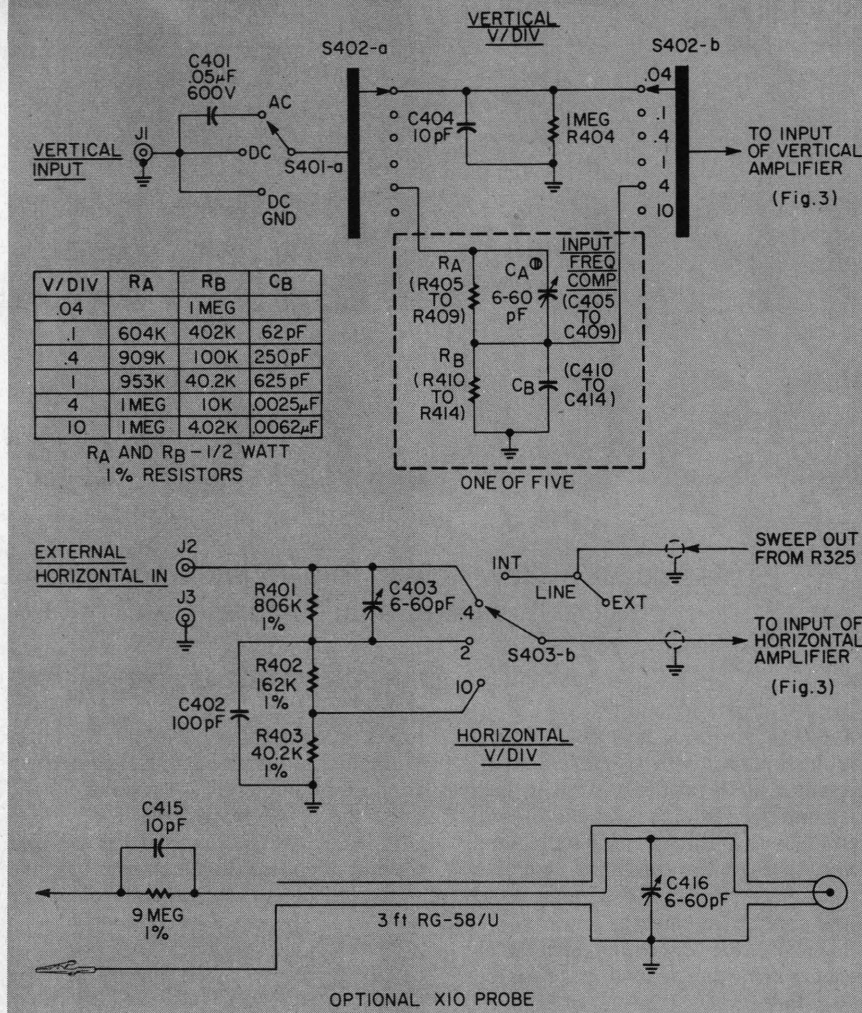
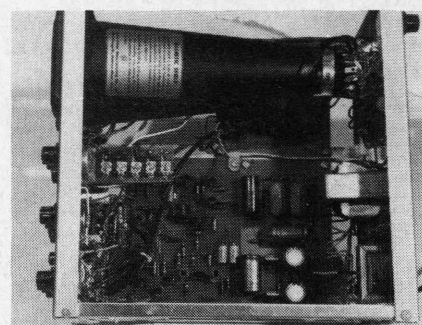


FIG. 2.—THE ATTENUATORS. Components for the vertical attenuator are mounted on a special circuit board. Also shown in the diagram of the optional multiplier probe.

sions at 0.04 volt-per-division) or a minimum of 0.01 volt (1 division at 0.01 volt-per-division). Capacitors C_A and C_B swamp out stray capacitances to keep the reactive division ratio exactly equal to the resistive division ratio at high frequencies. A1-4-10 step-sequence permits coverage of the 10-mV to 10-volt-per-division range with two poles of a standard six-position switch.

Vertical amplifier: The overall gain of the vertical amplifier (Fig. 3) is about 2000 in the full-gain ($\times 4$) position, and about 500 in the calibrated ($\times 1$) position of the vertical variable control. Resistor R201 and D201 provide input protection in the event of accidental overload. Source-follower Q201 and common-base amplifier Q202 form a trigger amplifier with a non-inverting AC gain of about 40 and a high input impedance.

Transistors Q203 and Q204 are switched on alternately by the zero-baseline flip-flop (Q307 and Q308, Fig. 4), connecting the base of source-follower Q205 alternately to the signal input and to ground. The stray capacitance of these FET's amounts to about 10 pF, and produces switching transients of about 10 μ S duration on the 1-megohm input line.



TOP VIEW of the scope. The amplifier board is beneath the CRT. The power-supply board is at the rear near the transformers mounted on the back panel. The sweep board is up front near the controls. The attenuator board, with its five trimmers, is on a bracket held by the vertical-sensitivity control. Astigmatism control is on rear panel near base of the CRT.

The switching frequency must therefore be held below a few hundred hertz to prevent those transients from being frequent enough to be seen on the CRT display.

Transistors Q206 and Q207 are wired as a variable-gain differential amplifier. Potentiometer R213 is the 10-mV calibrator and sets the gain to four times the indicated vertical sensitivity with R214

PARTS LIST (Attenuators, Fig. 2)

Resistors 1% tolerance or better, 1/2 watt

R401—806,000 ohms
R402—162,000 ohms
R403, R412—40,200 ohms
R404, R408, R409—1 megohm
R405—604,000 ohms
R406—909,000 ohms
R407—953,000 ohms
R410—402,000 ohms
R411—100,000 ohms
R413—10,000 ohms
R414—4,020 ohms
R415*—9 megohms

Capacitors

C401—.05 μ F, 600 volts, ceramic
C402—100 pF, Mylar
C403, C405—C414, C416*—6-60 pF ceramic trimmer
C404, C415*—10 pF, ceramic
C410—62 pF, mica
C411—250 pF, mica
C412—620 pF, mica
C413—.0022 μ F, Mylar
C414—.0062 μ F, Mylar

S401—miniature double-pole 3-position toggle switch (Alco MST205T)
S402—3-pole, 6-position rotary wafer switch
R403—2-pole, 6-position rotary wafer switch

Miscellaneous: printed circuit board

*Note: Components required for optional $\times 10$ probe

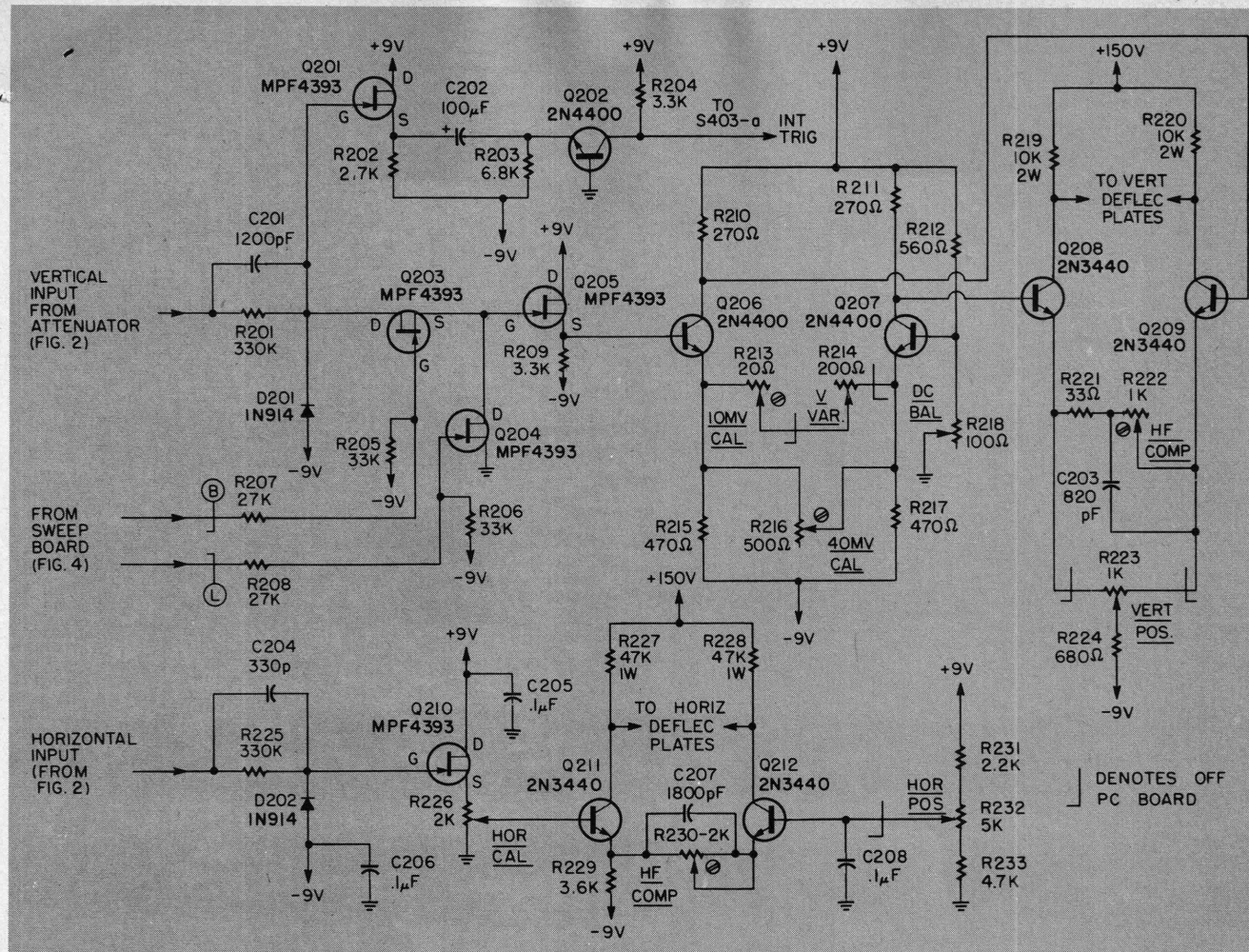
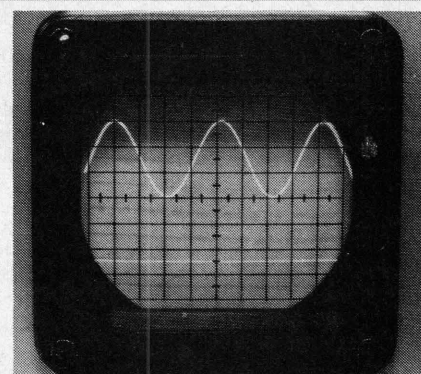


FIG. 3—SCHEMATIC DIAGRAMS of the vertical and horizontal deflection amplifiers. The latter is comparatively simple because its response is limited to the sweep frequencies.

at minimum resistance. Pot R216 is the 40-mV calibrator. It adjusts the indicated sensitivity with R214 at maximum resistance. Pot R218 is the DC balance control; it sets zero voltage between the two emitters at zero input in order that the gain control will not shift the vertical position.

Transistors Q208 and Q209 provide a

second stage of amplification, producing a maximum differential output of about 180 volts P-P. Capacitor C203 lowers the impedance between the emitters to track the decrease in impedance between the collectors caused by CRT plate capacitance at high frequencies. Since gain is essentially the ratio of those impedances, C203 tends to preserve



ZERO-BASELINE DISPLAY permits reading the DC component of this waveform. Scale factor is 1 V/div and the sinewave is 3 volts peak-to-peak riding on a 4-volt DC level.

a constant gain as frequency increases. Because an 820-pF trimmer would be large and unstable, we adjust the associated resistor (R202) to suit the capacitor, instead of vice-versa. Capacitor C203 thus determines the stage gain, and should be altered if necessary to produce a stage gain of about 50.

Horizontal amplifier: This amplifier (Fig. 3) is similar to the vertical amplifier except that the low-voltage differential stage is omitted and the entire gain (about 70) is achieved in the high-voltage stage. The differential output voltage required is about 250 volts P-P because the second (less sensitive) set of CRT deflection plates is used. Bandwidth is about 500 kHz.

PARTS LIST (Amplifiers, Fig. 3)

Resistors ½ watt, 10%, carbon composition, unless otherwise noted

R201, R225—330,000 ohms
R202—2700 ohms
R203—6800 ohms
R204, R209—3300 ohms
R205, R206—33,000 ohms
R207, R208—27,000 ohms
R210, R211—270 ohms
R212—560 ohms
R213—20 ohms, trimmer, vertical mount
R214—200 ohms, potentiometer
R215, R217—470 ohms
R216—500 ohms, trimmer, vertical mount
R218—100 ohms, trimmer, vertical mount
R219, R220—10,000 ohms, 2 watts
R221—33 ohms
R222—1000 ohms, trimmer, vertical mount
R223—1000 ohms, potentiometer
R224—680 ohms
R226, R230—2000 ohms, trimmer, vertical mount
R227, R228—47,000 ohms, 1 watt

R229—3600 ohms
R231—2200 ohms
R232—5000 ohms, potentiometer
R233—4700 ohms

Capacitors

C201—1200 pF, Mylar
C202—100 μ F, 15 volts, radial-lead electrolytic
C203—820 pF, mica
C204—330 pF, mica
C205, C206, C208—0.1 μ F, ceramic disc
C207, 1800 pF, Mylar

Semiconductors

D201, D202—1N914 or similar silicon diode
Q201, Q203, Q204, Q205, Q210—MPF4393 or similar N-channel FET (Motorola)
Q202—2N4402 or similar
Q206, Q207—2N4400 or similar
Q208, Q209, Q211, Q212—2N3440 or similar

Miscellaneous: PC or perforated circuit board, hookup wire, mounting hardware, transistor sockets, etc.

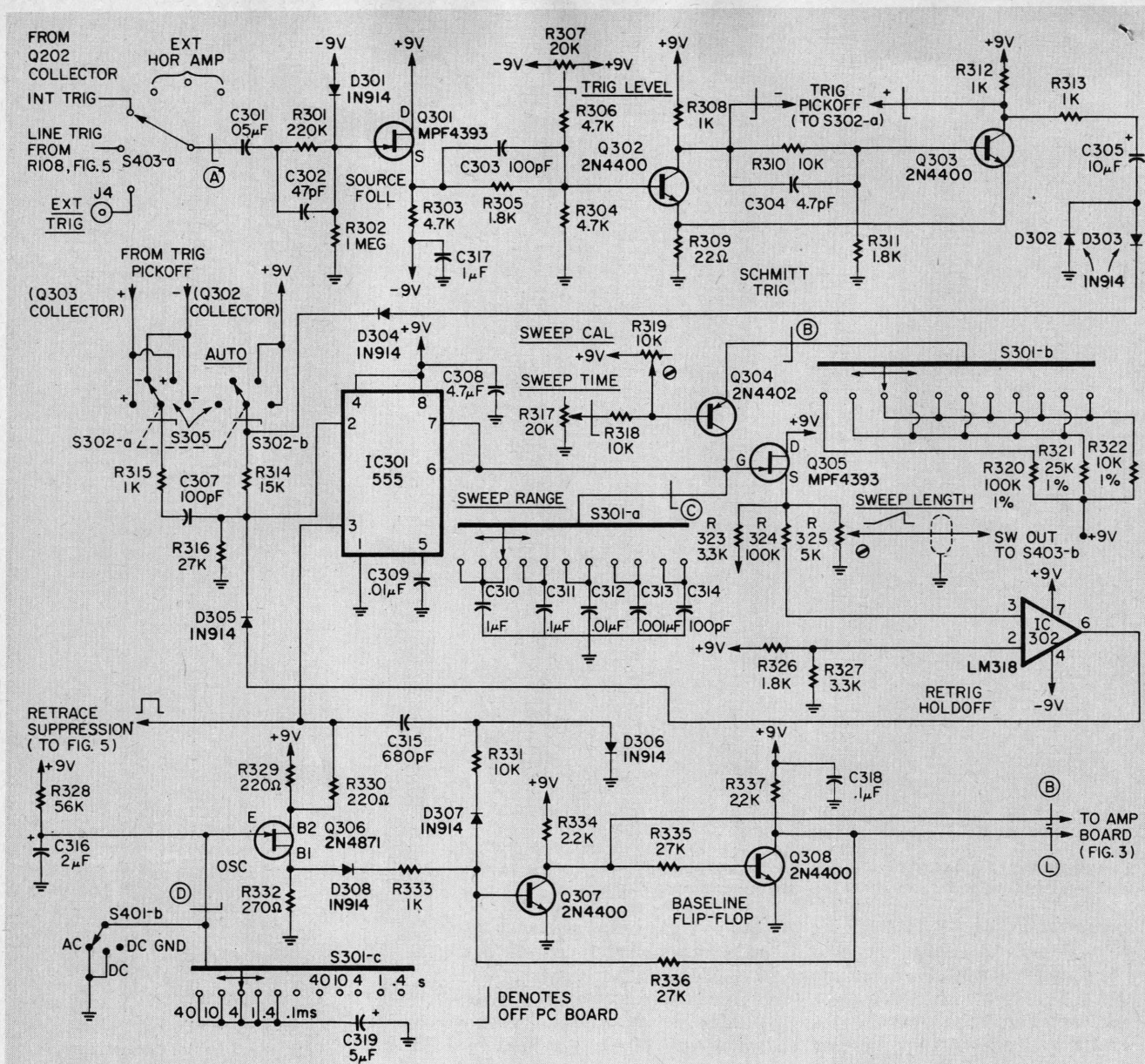


FIG. 4—THE SWEEP and zero-baseline generator circuits comprise the most complex sections of the instrument.

PARTS LIST (Sweep and zero-baseline generators, Fig. 4)

Resistors ½ watt, 10% unless otherwise noted

R301—220,000 ohms
 R302—1 megohm
 R303, R304, R306—4700 ohms
 R305, R311—1800 ohms
 R307, R317—20,000 ohms, potentiometer
 R308, R312, R313, R315, R333—1000 ohms
 R309—22 ohms
 R310, R318, R331—10,000 ohms
 R314—15,000 ohms
 R316, R335, R336—27,000 ohms
 R319—10,000 ohms, trimmer, vertical mount
 R320—100,000 ohms, 1%
 R321—25,000 ohms, 1%
 R322—10,000 ohms, 1%
 R323, R327—3300 ohms
 R324—100,000 ohms
 R325—5000 ohms, trimmer, vertical mount
 R326—18,000 ohms
 R328—56,000 ohms

R329, R330—220 ohms

R332—270 ohms

R334, R337—2000 ohms

Capacitors

C301—.05 μ F, 600 volts, ceramic disc
 C302, C304—47 pF, ceramic disc
 C303, C307—100 pF, ceramic disc
 C305, C306—10 μ F, 25 volts, axial-lead electrolytic
 C308—4.7 μ F, 25 volts, axial-lead electrolytic
 C309—.01 μ F, ceramic disc
 C310*—1 μ F, Mylar
 C311*—0.1 μ F, Mylar
 C312*—.01 μ F, Mylar
 C313*—.001 μ F, Mylar
 C314**—100 pF, ceramic trimmer
 C315—680 pF, ceramic disc
 C316—2 μ F, 25 volts, axial-lead electrolytic
 C317, C318—0.1 μ F, ceramic disc
 C319—5 μ F, 25 volts, axial-lead electrolytic

*Note: select to keep ratios within $\pm 1\%$

**Note: In prototype, C314 was made by connecting a 47-pF disc in parallel with a 6–60-pF ceramic trimmer

Semiconductors

D301—D308—1N914 or similar silicon diode
 IC301—555 timer
 IC302—LM318 op-amp (National)
 Q301, Q305—MPF4393 or similar N-channel FET (Motorola)
 Q302, Q303, Q307, Q308—2N4400 or similar
 Q304—2N4402 or similar
 Q306—2N4871 or similar unijunction transistor
 S401, S403—see attenuator parts list
 S301—3-pole, 11-position rotary wafer switch (Centralab PA-10009 or equal)
 S302—2-pole, 4-position rotary switch
Miscellaneous: PC or perforated circuit board, shielded cable, transistor and IC sockets, mounting hardware, knobs, etc.

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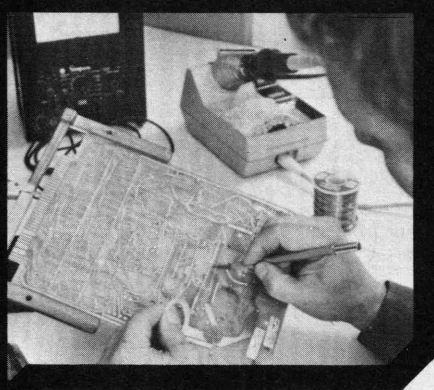
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OSCILLOSCOPE

continued from page 43

The LM318 op-amp (IC302) is used as a voltage comparator, holding the trigger input of the 555 positive until the timing capacitor has completely discharged. Premature triggering during retrace is thus prevented. The 555 provides a squarewave output at pin 3 that goes to +9 volts while the ramp is rising and drops to ground during retrace and hold-off. That line is capacitively coupled to the CRT grid to suppress the beam except during the sweep.

Baseline generator: The retrace-suppression line is used via R330 to synchronize the zero-baseline oscillator Q306, insuring that the switch from baseline to signal display will always occur at the start of a retrace when the beam is suppressed. For the lower sweep speeds, synchronization requires a slower oscillator, and for that C319 is switched in.

Each time unijunction transistor Q306 fires, C316 discharges through R332, setting flip-flop Q307-Q308 through D308 and initiating a baseline sweep. After the baseline sweep (or several sweeps if C316 is not discharged after the first one) pin 3 of the 555 goes low, bringing the base of Q307 low through C315 and D307, thus resetting the flip-flop for a series of signal displays.

Power supplies: The power supplies (Fig. 5) are entirely conventional except for the -900-volt tripler. Diodes D105 and D106 charge C102 to the peak negative voltage of the transformer secondary on the negative half cycle. On the positive half-cycle, C102 and the secondary appear in series to charge C101 to twice the peak secondary voltage (negative on top), through D103 and D104. On the next negative half-cycle, C101 and the secondary appear in series to charge C103 to three times the peak secondary voltage through D101 and D102. The drain on that supply is about 200 uA, so the 0.1 µF Mylar filters are quite adequate. Some of those capacitors are used at 20% or so above their rated voltage, but many have been tested at four times rated voltage with no breakdowns. Any string of five to ten Zener diodes adding up to about 900 volts will do for D109 through D113 if 180-volt Zener diodes are hard to find. Capacitor C105 filters out the 60-Hz noise picked up from the power transformer by the CRT heater winding.

We must breakoff our discussion of the oscilloscope's power supplies now and will conclude it next month when we will also go into construction, checkout and calibration. **R-E**



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Enjoy video programming throughout your house with the Video Master video distribution system.

RUDOLF F. GRAF and WILLIAM SHEETS

EXOTIC HOME-ENTERTAINMENT containing VCR's, satellite receivers, laserdisc players, video games, security cameras, and more are certainly enjoyable. However, they often fall short in convenience. The Video Master adds that missing convenience; it lets you set up your own video network and watch any video source on any TV in your home.

The Video Master, described in this article, consists of a series of converters that place all your video sources on unused UHF channels and then combines them with normal TV channels (terrestrial or cable) into one cable. That one cable

can then feed several TV sets for whole-house coverage. The desired video source is selected with the TV set's tuner. All of the TV's remote-control features are retained, and you'll be able to switch between any of the sources simply by changing the channel!

Keep in mind that it's illegal to insert unauthorized signals into any public cable system. An isolation amplifier, incorporated in the Video Master's design, prevents signals from feeding back into the cable system. *Do not omit the amplifier portion of the system for any reason.*

Figure 1 shows a block diagram of a typical system setup with the Video Master. It shows a TV antenna (or cable system), satellite TV receiver, VCR, security camera, video game, and laserdisc player, all feeding the Video Master. Notice that all inputs (except cable or antenna) are on Channel 3. There are five upconverters (1 through 5) and a buffer amplifier/power supply module, which supplies +12-volts DC to the upconverters.

The six outputs are combined with a splitter in reverse so that it acts as a signal combiner. The splitter feeds the combined output to all the TV sets in the sys-

tem. The combined output contains all terrestrial or cable channels (VHF 2–13, MID-BAND, UHF, etc.) in the original locations, plus five new channels—36, 39, 42, 59, and 61. Those five new channels carry the signals from the satellite TV, VCR, security camera, video game, and laserdisc player. Any TV on the system can select any signal source by tuning to its new channel.

A buffer amplifier and power supply module inputs the broadcast or cable channels into the system and prevents

signals from being fed back to the antenna or cable system. The amplifier also provides about a 7-dB gain to overcome unavoidable losses in the combiner network at the output of the system. The power supply is fed by either 12–14 volts AC or 15–20 volts DC; it supplies +12 volts DC at up to 180 milliamperes to power as many as five upconverters. (The amplifier/power supply module requires about 30 milliamperes, and each upconverter module requires about 35 milliamperes.)

Block diagrams of the upcon-

verter and amplifier modules are shown in Fig. 2. The upconverters consist of a preamplifier operating at 60–66 MHz, with 23-dB gain, and a two-pole bandpass filter between the amplifier and mixer. A double-balanced mixer combines the 60–66 MHz (Channel 3) TV signal with a local oscillator (LO). The LO is set to operate at a frequency 60 MHz below the low end of the desired output signal. For example, to obtain an output on Channel 39 (620–626 MHz), the LO must be at 560 MHz.

Several outputs appear at the mixer output: the original Channel 3 signal, the LO signal, and their sum and difference frequencies. A three-pole bandpass filter selects the desired output (the sum of LO + Channel 3) and rejects the difference frequency. (Theoretically, a double-balanced mixer produces no LO or Channel 3 output, but the mixer used here is not a perfect device, and those frequencies still appear.) The Channel 3 signal is severely attenuated, but the LO signal is suppressed by only 25 dB. Because the LO level must be about +7 dBm (decibels above one milliwatt) or about 0.3 to 0.5 volts rms at the mixer input, the LO appearing at the mixer output is still –18 dBm (about 30 millivolts rms). The desired Channel 39 signal is at about –34 dBm (approximately 4 millivolts rms). Therefore, the LO signal is about 16dB stronger than the desired Channel 39 signal, even using a balanced mixer and keeping the input signal level as high as possible (10 millivolts) to avoid generating excessive spurious signals. The LO signal can interfere with another channel 60 to 66 MHz lower (in this case, Channel 28).

If the LO signal is suppressed to less than half a millivolt, it causes no problems as long as it is placed outside an existing UHF channel. Because typically ten or fewer UHF channels can be received in any given area, ten or fewer new channels—or upconverters—will be needed. The upconverter channel outputs must be selected so as to

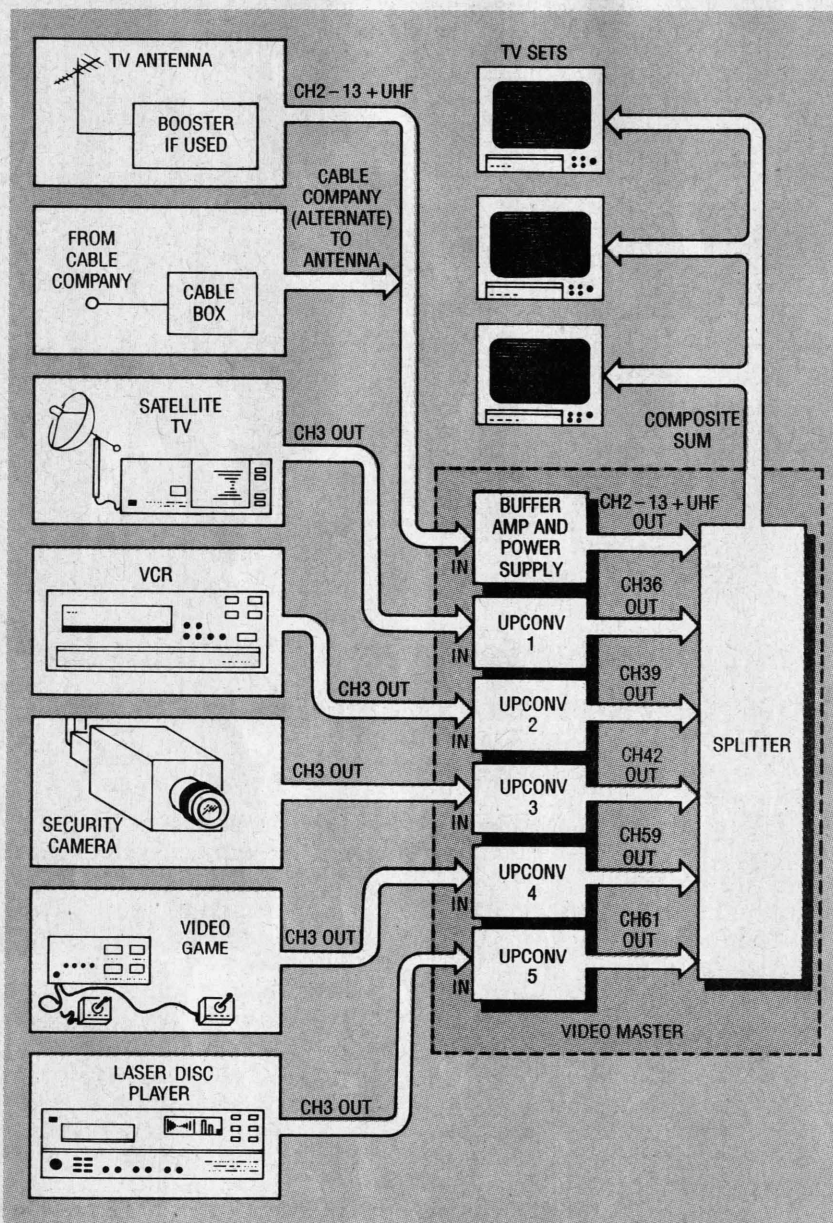


FIG. 1—THE VIDEO MASTER lets you convert Channel-3 outputs from many video devices to unused cable channels, combine them with your normal channels, and view them from any cable-ready set in your house, simply by tuning to the proper channel.

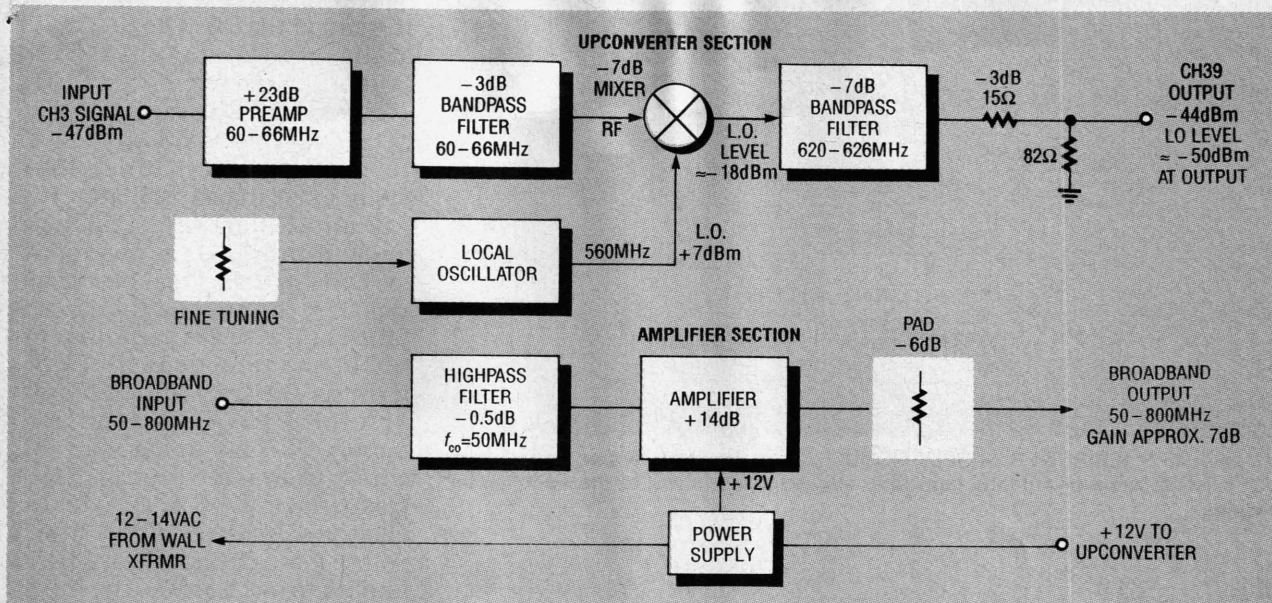


FIG. 2—BLOCK DIAGRAMS of an upconverter module and an amplifier/power supply module.

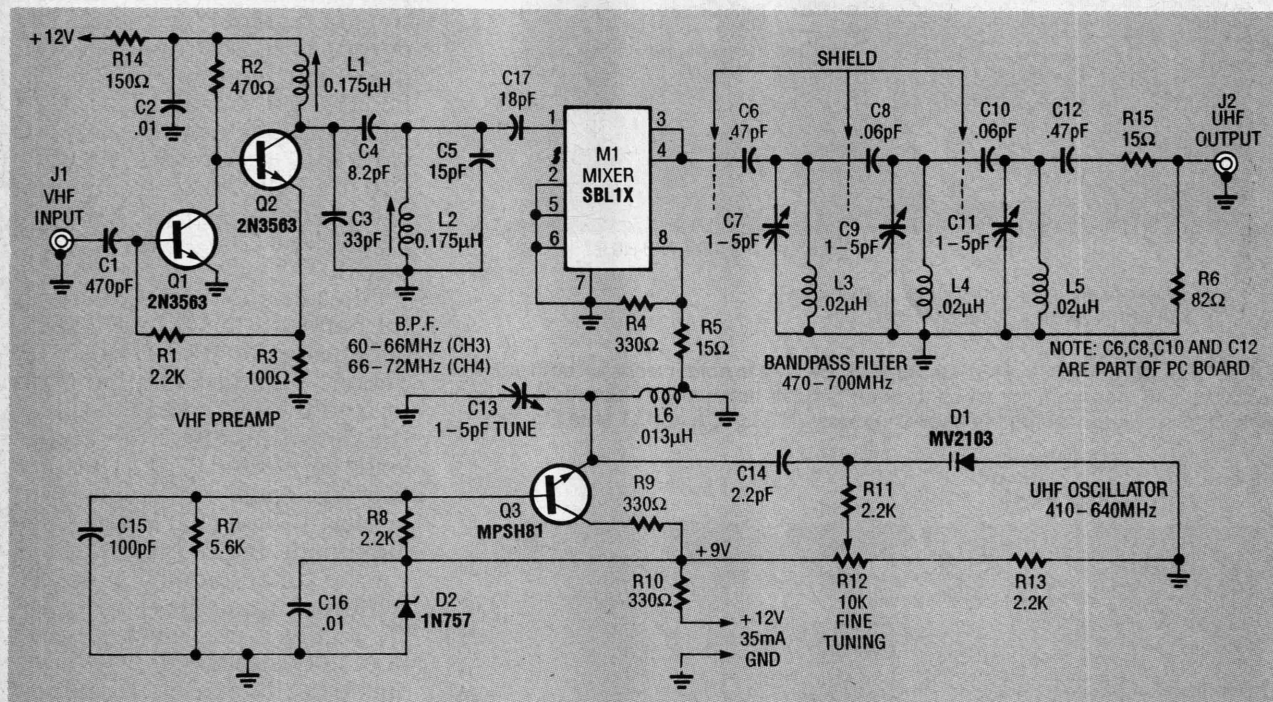


FIG. 3—VIDEO MASTER SCHEMATIC. This circuit inputs a VHF signal at J1 and converts and outputs it as a UHF signal at J2.

avoid placing the LO signal on top of an existing UHF channel. In our example, a Channel 39 upconverter output would have its LO at 560 MHz, and therefore it could interfere with Channel 28 and Channel 29. In an area where a UHF station exists on those channels, the upconverter should be moved up to Channel 41 or down to Channel 37 to avoid interference with

Channel 28 or Channel 29. In general, do not select an output frequency 10 or 11 channels higher than our existing UHF channel or any UHF channel to be used by another upconverter in the system.

To reduce stray signal pickup and interference problems in general, individual upconverter and amplifier/power supply modules are used. That elimi-

nates crosstalk problems, and simplifies shielding and circuit layout. The modular approach also lets you build only what you need, yet still allows future upgrades and expansion.

Circuitry

A schematic diagram of the upconverter is shown in Fig. 3, and its companion amplifier and power supply is shown in

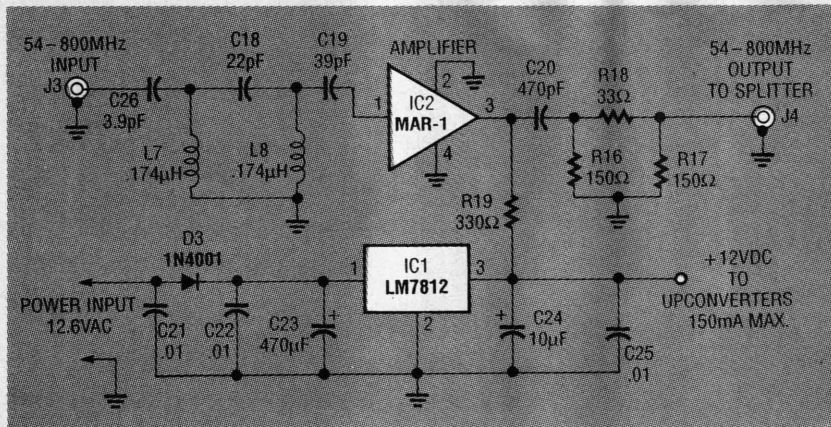


FIG. 4—AMPLIFIER/POWER SUPPLY MODULE. This circuit buffers and conditions your existing cable or antenna input and also supplies power to the upconverter modules.

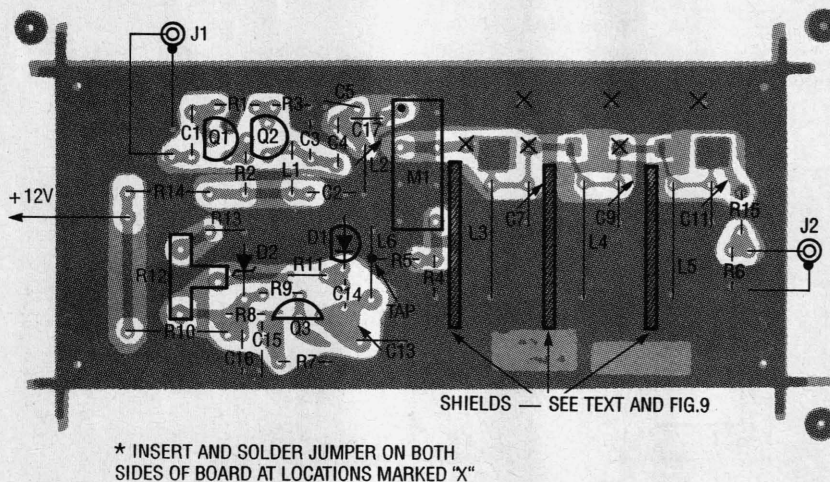


FIG. 5—UPCONVERTER PARTS PLACEMENT. Be sure to solder leads on both sides of the board, and insert jumper wires through unused holes and solder them on both sides. Don't forget to install the shields between M1, L3, L4, and L5 (see Fig. 9).

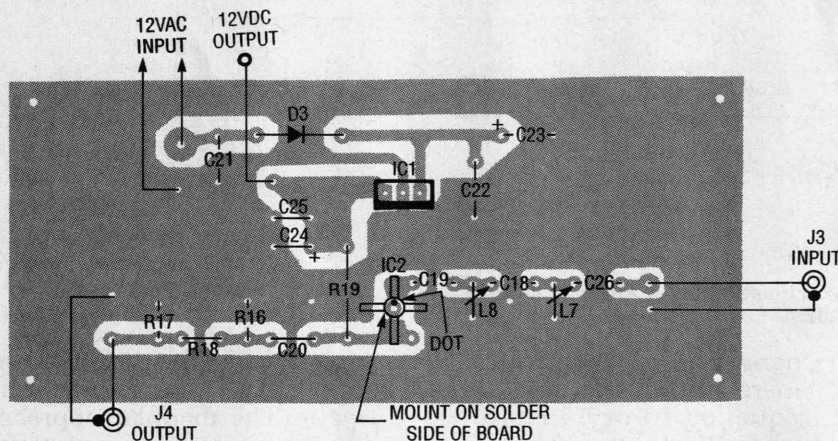


FIG. 6—PARTS-PLACEMENT DIAGRAM for the amplifier/power supply board.

Fig. 4. Channel 3 (VHF) input from a video device such as a VCR is supplied to J1. The signal level is assumed to be -47 dBm (about 1 to 2 millivolts). Capacitor C1 couples the input

to an amplifier consisting of Q1 and Q2. The collector of Q2 feeds a filter with a 60–66 MHz passband. The filter components are L1, C3, coupling capacitor C4, L2, C5, and match-

ing capacitor C6. The total gain at the mixer input (pin 1 of M1) is about 20dB referenced to J1.

Mixer M1 is driven by a UHF LO signal that is 60 MHz lower than the desired channel. Transistor Q3 is in the oscillator; R7, R8, and R9 are bias resistors, and C15 grounds the base of Q3 for UHF signals. Variable capacitor C13, coil L6, and the capacitance of the series combination of C14 and varactor diode D1 determine the frequency. Potentiometer R12, R13, and R11 supply DC bias to varactor D1, which allows fine tuning of Q1's frequency by ± 3 MHz. The oscillator signal (about 0.3 to 0.5 volts) is supplied through R5 and R4 to pin 8 of M1. The mixer output appears at pins 3 and 4 of M1, where about 4 millivolts of desired signal (the output) is present, along with 25–30 millivolts of residual LO signal. The output is fed to tunable band-pass filter made up of C6, C7, L3, C8, C9, L4 and C10, C11, C12 and L5. (Due to the very low values of capacitors C6, C8, C10, and C12, they are not discrete components, but are formed by traces on the PC board.)

Filter loss is about 7dB, and the bandwidth is about 10 MHz, depending on the center frequency. A simple attenuator pad formed by R15 and R6 reduces the detuning effect of varying loads connected to J2. The filter is a three-pole zero-ripple (Butterworth) type that allows easy alignment. In practice, the filter can be tuned simply by watching the output signal on a UHF TV receiver. It provides up to 50dB LO suppression with respect to the center frequency. Overall gain from J1 to J2 is about $+3$ to $+6$ dB. That allows for loss in combining the output of J2 with the outputs of additional converters.

Figure 4, the amplifier-section schematic, shows that the antenna or CATV input is applied to J3. A high-pass filter formed by C26, C18, C19, L7, and L8 attenuates unwanted signals (such as shortwave, CB, amateur, and AM) below 50 MHz. A monolithic microwave integrated circuit (MMIC) am-

plifier, IC2, has a broadband gain of about 14dB. Resistor R19 provides DC bias to IC2, and C20 couples the amplified output to resistors R16, R17, and R18 (which sets the total gain to about +7 to 8dB) and to J4. The amplifier compensates for the inevitable loss in the signal-combining network connected to J4.

A 12.6-volt AC wall-mounted transformer feeds components C21, D3, C22, and C23. Those components supply approximately 16-volts DC to the input of an LM7812 regulator (IC1), which supplies 12-volts DC to the rest of the circuit. (If desired, +15 to +20 volts DC can also be introduced to the supply. In that case, D3 would guard against reversed DC power-input polarity.)

Construction

The PC board for the upconverter is double-sided, and the board for the amplifier and power supply is single-sided; foil patterns are provided for both. Parts-placement diagrams for the two boards are shown in Figs. 5 and 6, respectively. Do not change the PC layout, because filter characteristics are dependent on it. PC boards and complete parts kits are available from the source given in the Parts List.

Although the upconverter boards are double-sided, they are not through-hole plated. Therefore, solder all component leads on both sides of the board wherever there is copper foil on both sides. In addition, you must place grounding jumpers in all holes marked with an "X," and solder them on both sides of the board to connect the top and bottom foils. Short lead lengths are important in RF projects because long leads can act as antennas. Also, mount all components snugly against the circuit board and clip their leads close to the board.

Although it would be possible to integrate inductors L3, L4, L5, and L6 into the PC board, the resulting printed inductors would have Q values that are too low, and there could be stray coupling and shielding diffi-

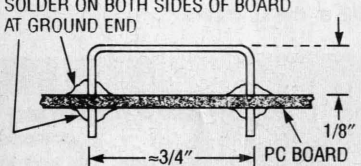
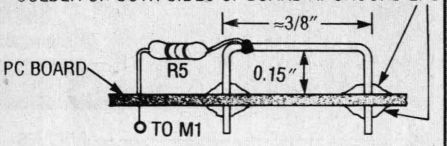
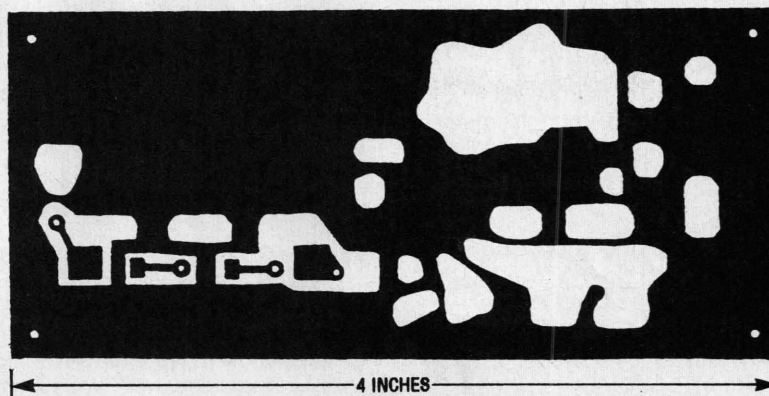
PART #	INDUCTANCE	CONSTRUCTION
L1,L2 65MHz FILTER COILS	.105-.230 μ H	 6 1/2 TURNS OF 22-GAUGE ENAMELED WIRE WIND WIRE IN THREADS OF 8-32 SCREW, FORM LEADS TO FIT PC BOARD, TIN LEADS, REMOVE SCREW, AND INSTALL SLUG
L3,L4,L5 UHF FILTER RESONATORS	0.02 μ H	18-GAUGE WIRE BENT TO FIT PC BOARD SOLDER ON BOTH SIDES OF BOARD AT GROUND END 
L6 OSCILLATOR INDUCTANCE	0.013 μ H	18-GAUGE WIRE BENT TO FIT PC BOARD MAKE TAP FOR R5 CLOSEST TO GROUND END OF L6. POSITION CAN BE ADJUSTED TO VARY L.O. DRIVE SOLDER ON BOTH SIDES OF BOARD AT GROUND END 
L7,L8 HIGH-PASS FILTER COILS AMP/POWER SUPPLY	0.174 μ H	 9-1/2 TURNS OF 22-GAUGE ENAMELED WIRE WIND WIRE IN THREADS OF 8-32 SCREW, FORM LEADS TO FIT PC BOARD, TIN LEADS, AND REMOVE SCREW (NO SLUG USED)

FIG. 7—COILS L1-L7 are made according to these instructions.

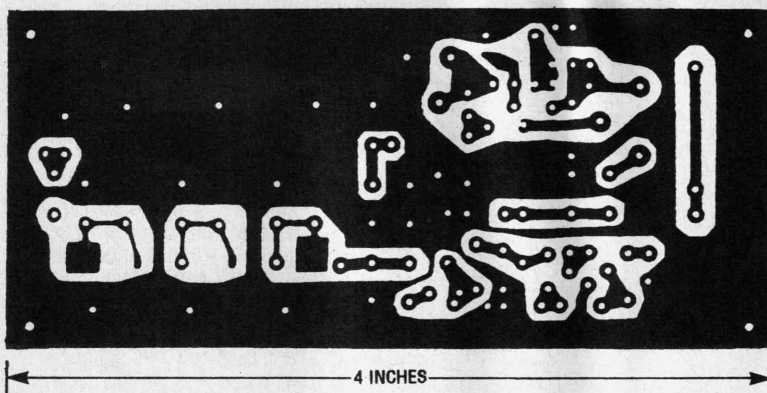


COMPONENT SIDE of the upconverter board.

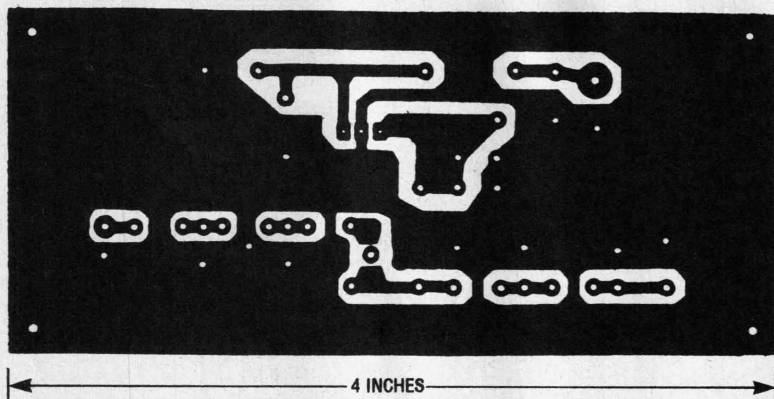
culties. A high value of Q is necessary for L3, L4, and L5 to achieve narrow filter bandwidth, and for L6 to stabilize the oscillator. Those coils are made from lengths of No. 18

AWG wire as shown in Fig. 7.

Note that shields—small scraps of G-10 double-sided PC-board material from 0.020- to 0.062-inch thick—must be soldered after standing them on



SOLDER SIDE of the upconverter board.



SOLDER SIDE of the amplifier/power supply board.

edge on the top ground plane of the upconverter PC board between M1, L3, L4, and L5 in the locations shown in Fig. 5. The shields are necessary for proper filter performance because they keep down spurious outputs—especially the LO residual leakage. The shields must be well soldered to the top of the PC board.

Mount each converter in its own enclosure. A suitable case is included with the previously mentioned upconverter kits. Suitable connectors are F, BNC, TNC, or SMA—do not use UHF or RCA connectors. Figure 8 shows a suitable packaging scheme for the upconverter board and the amplifier board. Do not omit the three shields in the upconverter filter section, as shown in the photo of the board in Fig. 9. Figure 10 is a photograph of the amplifier/power supply board.

Figure 11 shows one way to mount several modules together to make up a system. The module outputs all connect to the "outputs" of a passive split-

PARTS LIST

All resistors are 1/8 watt, 5%, unless otherwise noted.

R1, R8, R11, R13—2200 ohms
R2—470 ohms
R3—100 ohms
R4, R9—330 ohms
R5, R15—15 ohms
R6—82 ohms
R7—5600 ohms
R10, R19—330 ohms, 1/4-watt
R12—10,000 ohms, potentiometer
R14—150 ohms, 1/4-watt
R16, R17—150 ohms
R18—33 ohms

Capacitors

C1, C20—470 pF, ceramic disc
C2, C16, C21, C22, C25—0.01 μ F, ceramic disc
C3—33 pF, 5%, NPO
C4—8.2 pF, 5%, NPO
C5—15 pF, 5%, NPO
C6, C8, C10, C12—Part of PC board
C7, C9, C11, C13—1–5 pF trimmer
C14—2.2 pF $\pm$ 0.5 pF, NPO
C15—100 pF, 10%, NPO
C17—18 pF, 5%, NPO
C18—22 pF, 5%, NPO
C19, C26—39 pF, 5%, NPO

C23—470 μ F, 25 volts, electrolytic
C24—10 μ F, 16 volts, electrolytic

Semiconductors

IC1—LM7812 12-volt regulator
IC2—MAR1 MMIC
D1—MV2103 varactor diode
D2—1N757 9-volt Zener diode
D3—1N4001 diode
Q1, Q2—2N3563 NPN transistor
Q3—MPS H81 NPN transistor

Other components

L1, L2—6 1/2 turns No. 22 AWG enameled wire wound on 8-32 screw with 8-32 ferrite slug (see Fig. 7)
L3—L5—0.02 μ H (3/4-inch No. 18 AWG wire, see Fig. 7)
L6—0.013 μ H (1/2-inch No. 18 AWG wire, see Fig. 7)
L7, L8—0.174 μ H (9 1/2 turns No. 22 AWG enameled wire wound on 8-32 screw)
M1—SBL1X mixer
J1—J4—chassis-mount F-connector

Miscellaneous: PC boards, one shielded enclosure per module, hardware, feedthroughs or bushings for power inputs and outputs,

12.6-volt AC wall transformer, two feet of 22-gauge enameled copper wire, two feet of 18-gauge tinned copper wire, 2- or 4-way splitter, 75-ohm cable, solder

Note: The following items are available from North Country Radio, PO Box 53, Wykagyl Station, New Rochelle, NY 10804:

- Amplifier/power supply kit (includes case, hardware, connectors, PC board, and all parts)—\$29.50
- Upconverter kit (includes case, hardware, connectors, PC board, and all parts)—\$34.50
- Four-way splitter—\$5.50
- 12.6-volt AC wall transformer—\$ 9.50

Please add \$3.50 S&H (foreign orders \$5.00) to any order. NY residents must add sales tax. For a catalog of other kits, please send \$1 (refundable on next order) or SASE (52 cents postage) to North Country Radio at the above address.

ter. A passive splitter can be used as a combiner simply by running it backwards. That won't work for an active, amplified splitter.

The upconverter modules have a 3-dB gain, which is adequate for compensating for splitter loss. Because most video devices have an RF output of about 3dB above 1 millivolt, approximately 1 millivolt of UHF signal will appear at the system output for each channel, assuming the use of a four-way splitter and three upconverters. The cable level should be around 1 millivolt per channel, which will give about 50dB or better signal-to-noise ratio for the average TV receiver. Levels lower than 200 microvolts might yield a snowy picture. If necessary, the system output can be run through a distribution amplifier. Remember to terminate all unused splitter ports with 75-ohm terminating resistors.

Test and alignment

Alignment of the completed unit requires a video source on Channel 3 (your VCR will do) and a digitally tuned TV set. A frequency counter will also be helpful in this procedure.

First check out the amplifier/power supply. Connect a source of 12 to 14 volts AC, of at least 250 milliamperes to the junction of C21 and D3, and connect the remaining lead to ground—a plug-in wall transformer is recommended. Alternatively, a DC source of 15 to 20 volts can be used, with the positive lead to the C21-D3 junction, and negative lead to ground. Regardless of the supply you use, verify that there is 15 to 20 volts DC across C23.

Next check for +12 volts at the junction of IC1, C24, C25, and R15. If there is less than 11.5 volts or more than 12.6 volts, check to see if IC1 is defective or improperly inserted in the PC board. Check to be sure that IC1 does not get hot. If all tests are passed so far, check for +4 to +7 volts at pin 3 of IC2. Next check for infinite resistance from the center pin of J3 to ground. Next, check the re-

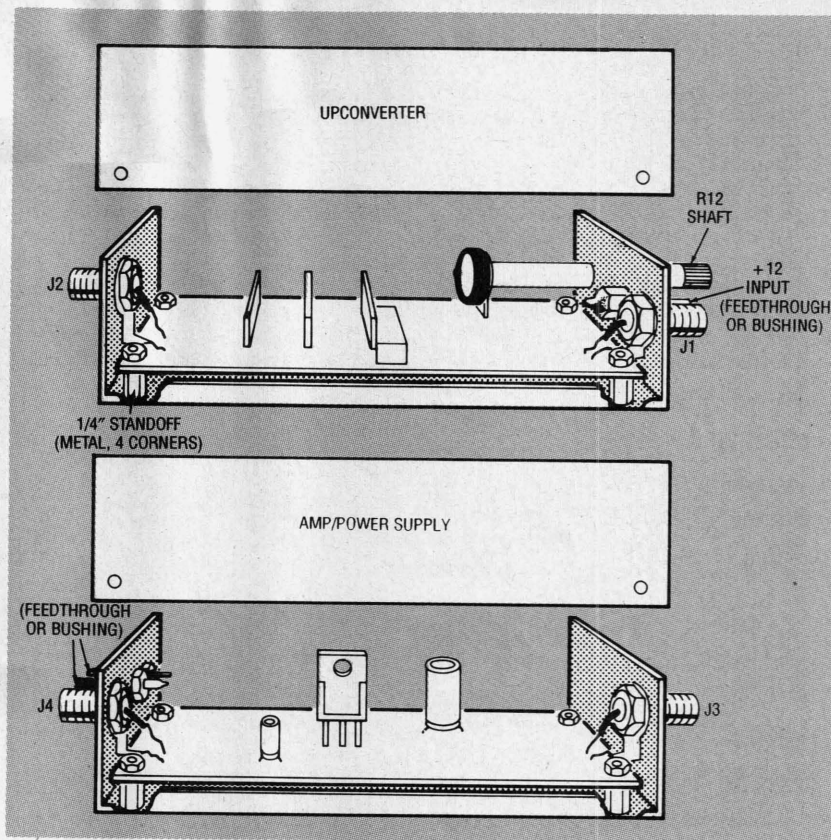


FIG. 8—SUGGESTED PACKAGING SCHEMES for the upconverter and amplifier/power supply boards.

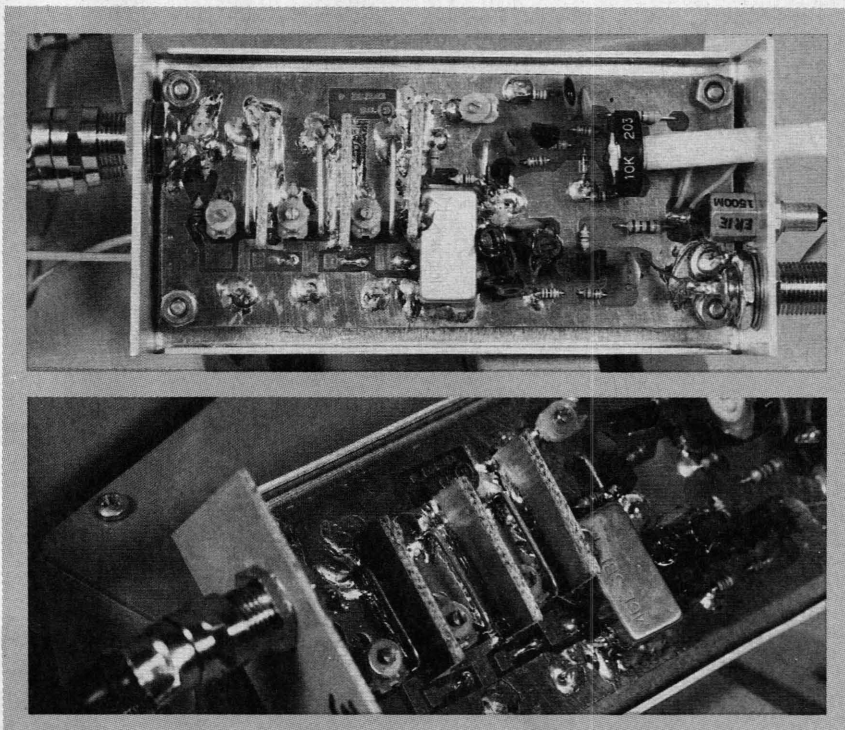


FIG. 9—UPCONVERTER BOARD. Note the shields, made from scraps of double-sided PC board material, that are placed between M1, L3, L4, and L5.

distance from J4's center conductor to ground; it should be about 80 ohms. Mount the

board in a case, apply power, and connect a TV receiver to J4, and the antenna to J3. Normal

CABLE BOXES

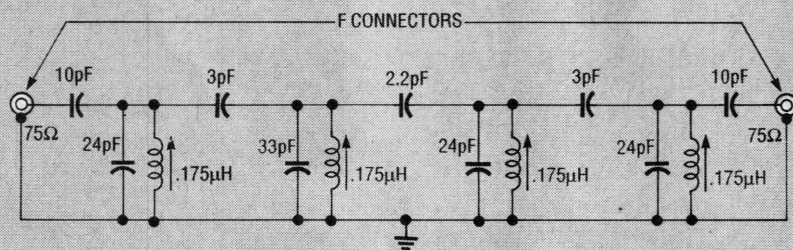
There are many different types of cable boxes, but as far as the Video Master is concerned, there are only two types. One type contains a built-in video modulator and its output is spectrally clean enough for the Video Master.

However, many cable boxes are simply RF converters, and there are other frequencies mixed in with the Channel-3 or -4 output. You can easily test this by changing channels on the TV's tuner. If the TV can receive any channels other than Channel 3 (or 4), you could have problems with the upconverter modules that will show up as lines, ripples, noise, and beats in the picture.

If you have any of those problems and you're sure that no stray signals are

leaking into your system, build the filter shown here. It's designed to pass only Channel 3, but it can be retuned for Channel 4. Its capacitors are all silver mica or NPO ceramic. The coils are $0.175\ \mu\text{H}$, and can be made in the same way as L1 and L2 in the upconverter modules.

The filter can be built on a scrap piece of G-10 copper-clad PC board material. It should be mounted in a shielded box and provided with F connectors for best results. It is aligned by peaking the coils for maximum signal on Channel 3. A VCR can be used as a signal source. The filter should be inserted between the cable box and the input connector on the appropriate upconverter. Ω



NOTE: FILTER IS SYMMETRICAL — INPUT AND OUTPUT ARE INTERCHANGEABLE

This filter will remove all signals except Channel 3 (or 4) from the output of your cable box.

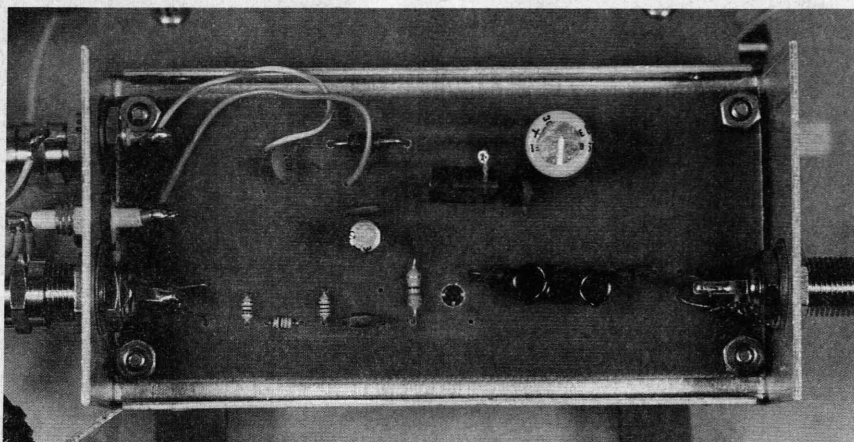


FIG. 10—AMPLIFIER/POWER SUPPLY. This board is mounted in the same type of case as the upconverter modules.

TV reception should result, with no loss of picture quality. If RF test equipment is available, measure the gain from J3 to J4. About 7 to 8dB should be obtained at 450 MHz (UHF), and slightly more on VHF (100 MHz). That completes the amplifier/power supply tests.

The upconverter board is tested as follows: After the board

has been visually checked for shorts, solder bridges, and correct component placement, install the board in its case. Connect a Channel-3 source to J1 and connect a TV receiver to J2. Use 75-ohm cable. Apply power to the 12-volt input (the junction of R10 and R14), and check for the following voltages:

- Junction of D2 and R10—

+8.4 to 9.5 volts

- Wiper of R12—+3 to 9 volts depending on setting of R12
- Junction of R11 and D1—+3 to 9 volts depending on setting of R12
- Emitter of Q3—+6 to 7.5 volts (adjusting C13 should vary the voltage by ± 0.1 volt—this verifies that Q3 is oscillating)
- Base of Q3—+6 to 7.5 volts (adjusting C13 should vary the voltage by ± 0.1 volt—this verifies that Q3 is oscillating)
- Emitter of Q2—+1 to 1.2 volts
- Base of Q2—+1.8 to 2.1 volts
- Collector of Q2—+8.5 volts (typical)
- Pin 8 of M1—0.3 to 0.5 volts RMS (this test is optional, and can only be done with an RF voltmeter)

Tune the TV receiver to the UHF channel on which you would like the upconverter to produce a signal. Set the slugs in L1 and L2 halfway in the coil winding. Set C7, C9, and C11 so that their plates are halfway engaged, and C13 fully engaged. Set R12 to mid-position. Turn on the source connected to J1. Slowly rotate C13 with a plastic alignment tool; at several points the TV set should exhibit a response of some kind. (If you have a frequency counter, connect it to pin 8 of M1 and set C13 for the correct oscillator frequency.) When you get a response, you might see a very weak picture, but at first you will probably only hear audio. Note the position of C13. Now look for other responses; the correct one will be where C13 is set at greater capacitance (more of the plates engaged).

The oscillator can produce an output either on the high or low side of the desired channel—you want the low side, otherwise the converter output will have picture and sound frequencies inverted from the usual positions. Next, slowly adjust C7, C9, and C11 for best picture and sound. Now go back to L1 and L2 and adjust for best picture quality as well as sound quality. Readjust C7, C9, and C11 for the best picture and sound. Repeat any alignment as needed.

The adjustment of trimmer

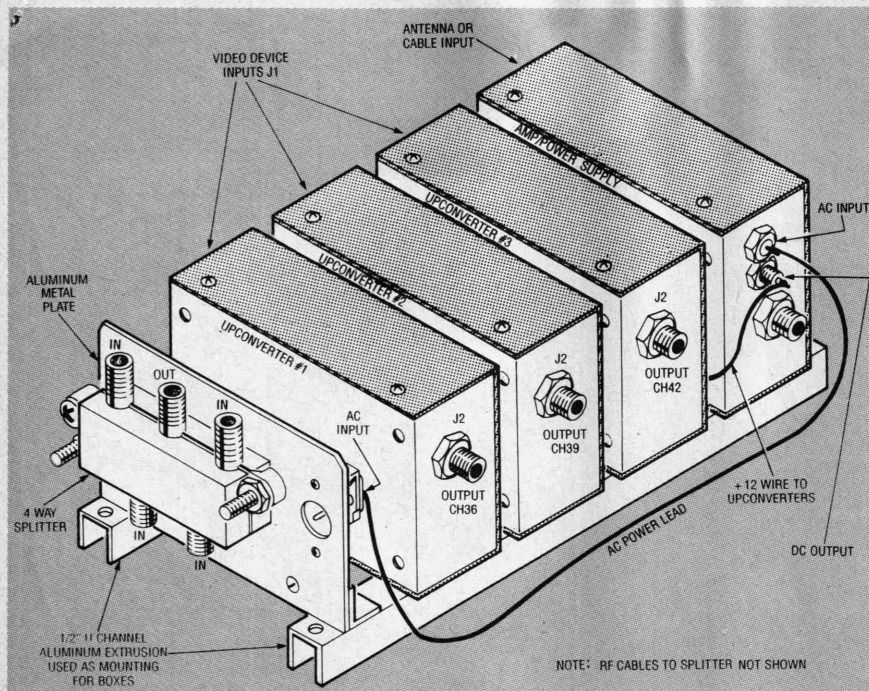


FIG. 11—ALL OF THE MODULES for the prototype Video Master are mounted together on a pair of aluminum rails, with everything feeding into a 4-way splitter mounted on one end of the assembly.

capacitors C7, C9, and C11 is very critical. Some difficulty

might be experienced at first, "getting in the ballpark," since

the bandpass filter is quite sharp (10 MHz), and it will have high attenuation when misaligned. Once you get a picture of any kind, the rest is easy. If the unit appears to work but the TV set tuning is critical, the picture "grainy," or the color poor, make sure that C13 is set on the proper side. (As mentioned before, there will be two settings, and the lower frequency is correct.) When aligned, the picture on the selected UHF channel should be of excellent cable quality.

After alignment is complete, verify the fine-tuning adjustment R12. Normally R12 is left in the halfway position, and adjusted only to touch up the frequency setting.

The upconverter can now be installed in your TV system, and should operate reliably with no attention from you. For overall system stability, we recommend leaving the upconverter system on all the time, hence no switch is installed in the system. □

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